NAVSEA OD 30393

SECOND REVISION

DESIGN PRINCIPLES AND PRACTICES FOR CONTROLLING HAZARDS OF ELECTROMAGNETIC RADIATION TO ORDNANCE

(HERO DESIGN GUIDE)

DISTRIBUTION STATEMENT A
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THIS PUBLICATION SUPERSEDES NAVSEA OD 30393, FIRST REVISION DATED 15 SEPTEMBER 1974

PUBLISHED BY DIRECTION OF COMMANDER, NAVAL SEA SYSTEMS COMMAND

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FOREWORD

1. NAVSEA OD 30393 has been prepared as a guide for HERO preventive techniques to be applied to the design and construction of weapon and subsystems. The information contained herein should not be construed as a specification but as an aid in implementing the requirements of MIL-STD-464, “Department of Defense Interface Standard: Electromagnetic Environmental Effects Requirements for Systems.”

2. This design guide is a compilation of data gathered from a number of sources. NAVSEA OD 30393, First Revision was used as the primary source, while other publications, in particular, NAVAIR AD 1115, “Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment,” Third Edition, June 1988, Naval Air Systems Command, were consulted extensively.

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CHAPTER 1

INTRODUCTION

1-1. BACKGROUND.

Modern communication and radar transmitters can produce high-intensity electromagnetic environments (EME's) that are hazardous to ordnance and to the attending personnel and associated equipment. These EME's can cause premature actuation of sensitive electrically initiated devices (EID's), classically known as EED's or electroexplosive devices. The EME can also damage or trigger solid-state circuits, damage or cause erratic readings in test sets, or produce sparks that can damage or burn out components of essential circuits of the ordnance system. If these electronic components are part of an EID's firing circuit, then electromagnetic (EM)-induced effects can result in the initiation of the EID. The trend of developing communication and radar transmitter/antenna systems with greater effective radiated power continues to augment the electromagnetic radiation hazards to ordnance problem. The nature and character of these electromagnetic fields are discussed in chapter 2.

This design guide is intended primarily to assist the ordnance system developer solve the problem of premature actuation or degradation of EID's. This requirement, therefore, places a responsibility on the designer to address the radio frequency (RF) susceptibility of the total ordnance system: the ordnance, its packaging, the test and checkout sets, any remote control interfaces, and any external prime power connections. Figure 1-1 illustrates the basic elements of a weapon and the typical interfaces with other elements of the missile system.

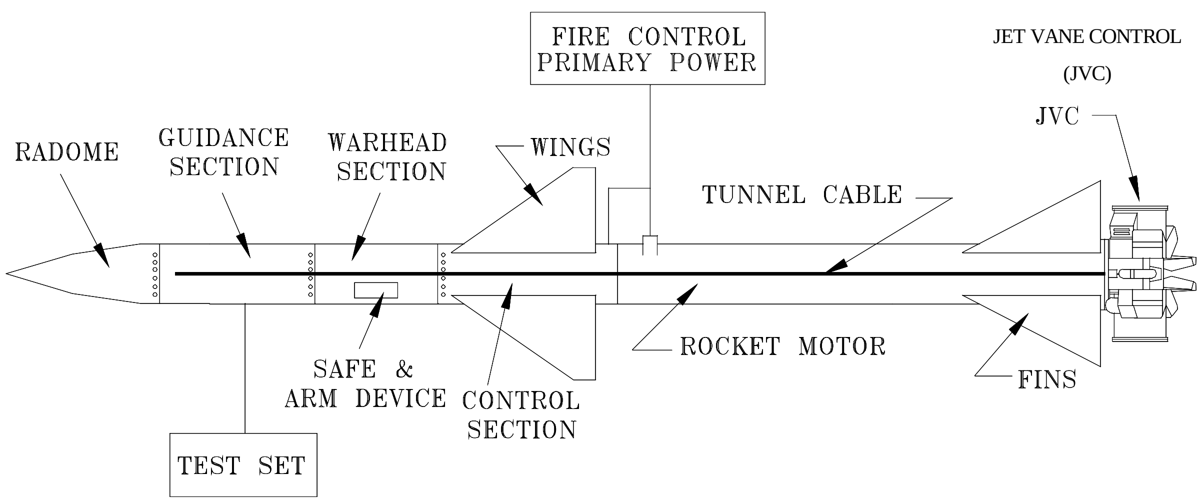


Figure 1-1. Functional Elements of a Missile System

The problem of premature actuation of EID's within the ordnance system as a result of exposure to the EME is known as Hazards of Electromagnetic Radiation to Ordnance (HERO). The solution to the HERO problem requires an understanding of the nature of the environment, the response of circuits and devices to radiated or conducted EM energy, and the methods and techniques used to mitigate the adverse effects of the EME on the ordnance system.

The advent of smart weapons and ordnance with logic and digital microcircuits makes weapons and ordnance more tactically effective yet compounds the radiation hazards to ordnance. The function of the EID's within the ordnance is to cause some explosive, pyrotechnic, or a mechanical output resulting from an explosive or pyrotechnic action, or have some dynamic mechanical, thermal, or electromagnetic output. Examples of EID's include hot bridgewire devices, conductive composition electric primers, carbon bridge devices, semiconductor bridge EED's, laser initiators, exploding foil initiators (EFI's), burnwires, and fusible links. These devices and their characteristics are discussed in chapter 3.

Energy from the EME can enter the ordnance item through discontinuities in its skin, such as ports, cracks and joints, and it can couple into circuits controlling the EID's. More energy will generally enter the ordnance item when the weapon ports are open than will enter when the ports are closed. The energy can also be conducted into the item by firing leads and other electrical conductors, such as test leads and power cables. In general, ordnance is more susceptible to EME during assembly, disassembly, handling, loading, and unloading than at any other time because personnel and tools are used at a time when the ordnance may be poorly grounded or improperly shielded. The attachment of external cable assemblies and test sets to an ordnance item will usually increase its susceptibility to RF energy. Chapter 4 provides a detailed discussion of typical firing system designs and their potential susceptibility to RF energy.

For most ordnance, a HERO problem is inevitable unless the designer recognizes the possible hazards and organizes all phases of the development so that the hazard is precluded in the initial design. Retrofitting after a HERO problem is discovered at some later stage of development is, at best, expensive and time consuming, and often detracts from the tactical reliability of the ordnance.

1-2. OBJECTIVES OF THIS DESIGN GUIDE.

This design guide is written to amplify and augment the Department of Defense (DoD) Interface Standard, MIL-STD-464, "Department of Defense Interface Standard: Electromagnetic Environmental Effects Requirements for Systems," and its objectives are to:

a. Define and describe the hazardous EME.

b. Provide ordnance system designers with engineering data sufficient to understand the HERO problem and to determine the solution to the problem.

c. Identify and recommend specific design and fabrication practices to preclude HERO.

While it is recognized that each system will be unique with respect to HERO, an effort has been made to present recommended design practices and associated engineering data and theory in a manner that will assist the designer in adapting the various recommendations to his particular situation.

1-3. POSSIBLE SOLUTIONS.

The resolution of the HERO problem might be approached logically in one of the following ways:

a. Eliminate all EID's from the ordnance. This approach would certainly resolve the HERO problem; however, EID's have unique characteristics that make them an essential part of the ordnance system. Therefore, eliminating EIDs altogether from ordnance is most likely not a viable solution.

b. Keep all ordnance containing EID's physically isolated from the EME. This may be an option for shore facilities where there is a lot of real estate to work with; however, it is usually impossible to keep the ordnance separated from the normal Naval EME, particularly on-board ship.

NOTE

> Currently, some Fleet weapons require physical separation from (or the silencing of radio or radar transmitters) during certain phases of the assembly, disassembly, handling, or loading. Current restrictions are published in NAVSEA OP 3565/NAVAIR 16-1-529 Volumes 1 and 2, which is available only to Fleet and shore stations. Restrictions are not desirable and are less acceptable to the Fleet as new and safe ordnance systems are developed to replace the ones that have restrictions.

c. Silence the transmitters generating the hazardous EME when ordnance containing EID's is present; see note above. However, it is usually not practical to silence the radio and radar transmitters since they are essential to normal Fleet operations. Also, there are situations and/or conditions in which part of the EME is not subject to Naval or DoD control.

d. Design the ordnance system to survive and operate as intended in its expected operational EME throughout its stockpile-to-safe separation sequence. While solutions a, b, and c would eliminate the HERO problem, they are not practical. Therefore, the only practical solution is to design every system in such a way that sufficient protection is afforded the EID's to eliminate the HERO problem under all conditions that may be encountered.

1-4. BASIC APPROACH.

There are several approaches that can be considered for solving the HERO problem. These approaches, discussed in the following paragraphs, use a number of techniques and methods to reduce the coupling of RF energy to acceptable levels at the EID. The designer can select the best approach that satisfies the requirements of the system.

1-4.1. CONTINUOUS RF SHIELD. The basic approach consists of enclosing all EID's and their firing circuits (including all power sources, transmission lines, and switching and arming devices) within a continuous electromagnetic interference (EMI) shield or "conductive box."

This approach requires that proposed design techniques and fabrication methods will ensure that the EME cannot penetrate into the shielded area. This concept is illustrated in figure 1-2 and requires that the integrity of the RF shield be designed and maintained throughout the life cycle of the system.

Maintaining the shielding integrity of an RF-tight enclosure requires that the mechanical design and fabrication techniques ensure that EM energy cannot couple into the shielded enclosure due to poor mating of parts, surfaces, or openings. The advent and use of non-conductive and/or composite materials complicates the total shielding approach and requires the implementation of conductive coatings and RF-tight mating surfaces at the non-conductive interfaces. Methods and techniques for providing effective and reliable shielding with appropriate bonding and ground are described in chapter 5 (Shielding), chapter 6 (Bonding), and chapter 7 (Grounding).

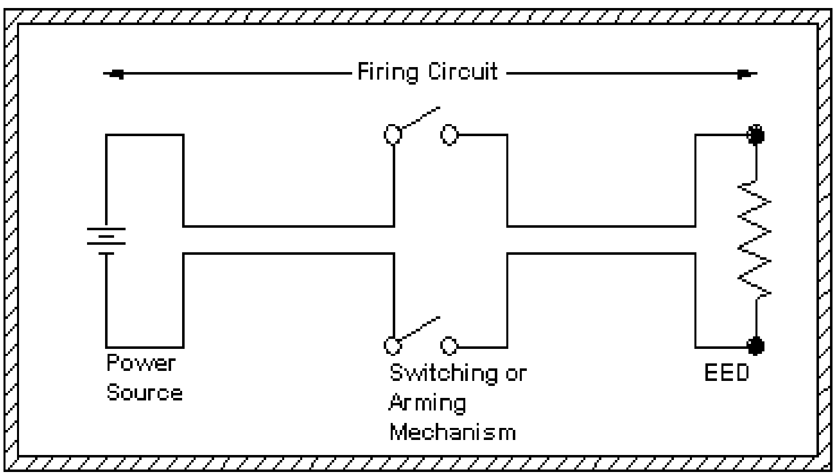


Figure 1-2. Conductive Box Concept

The "conductive box" concept may be possible for the ordnance itself, but often is not practical for the total ordnance system, which includes the installation, testing and checkout, and powering of the system while attached to some host platform. Under these conditions, an alternative to the totally conductive box must be considered; see paragraph 1-4.2.

1-4.2. SHIELDED COMPARTMENTS AND INTERCONNECTIONS. Another method to exclude RF energy from coupling into ordnance is to compartmentalize the system into shielded subsystems connected with RF-shielded or protected interconnects as shown in figure 1-3. This technique requires that the RF shielding integrity of each subsystem and of each interconnection be designed so that the EME cannot couple into the system at any point.

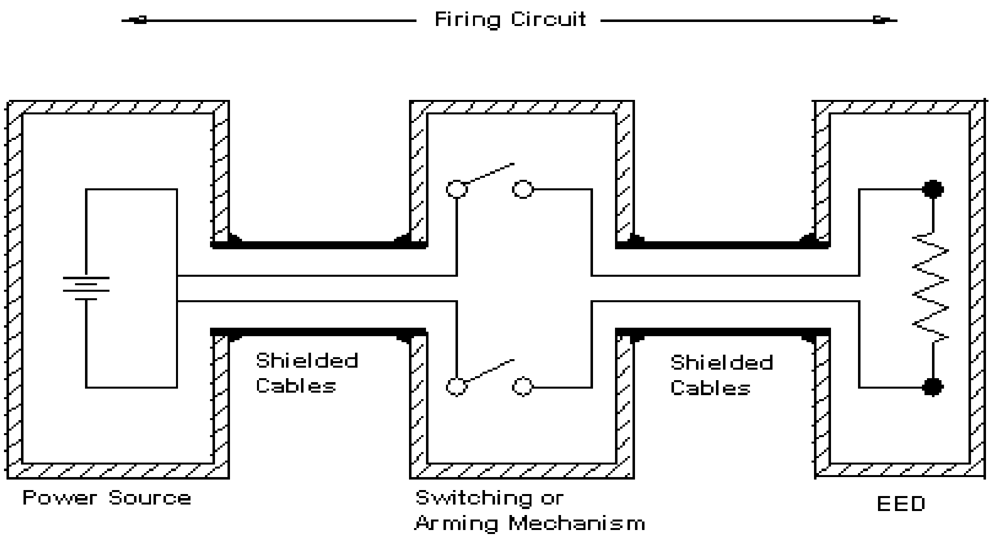


Figure 1-3. Shielded Compartmented Ordnance with Interconnect Protection

This approach requires care similar to the totally shielded, conductive box approach in shielding each of the individual compartments. It also requires the extensive use of shielded cables and/ or EMI type connectors for cable interconnects. Another means of connecting compartments that would eliminate RF coupling into the system via the cabling is to use non-conductive interfaces such as fiber optic cabling. The methods and techniques for the shielding and protection of all apertures and access panels, and the suppression of the EME on all penetrations into or out of the system, are described in chapters 5 through 7.

NOTE

> Major portions of chapters 5-8 are extracted from NAVAIR AD1115, "Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment," June 1988.

1-4.3. EMI FILTERING. Most ordnance require breaking electrical connections when the parts of the system are physically separated. Thus, it is often impossible or impractical to keep all conductors within one continuous shield. Therefore, EM energy must be excluded by some other method. It can be excluded from a shielded enclosure at a connector by means of an EMI filter (a low-pass filter). The filter is typically used to dissipate the EM energy instead of reflecting it at an impedance mismatch, as is usually the case. Because the generator and the load impedance are unknown and vary frequency, reflection due to mismatch of impedances cannot be relied on to protect the ordnance system. One precaution to be noted is that the heat generated in the filter by dissipation of the EM energy must be prevented from actuating the EID. This can be accomplished by providing a separation of the EID and the filter or by providing a heat sink. The proper use of a filter is illustrated in figure 1-4. For further details on EMI filtering techniques, refer to chapter 8.

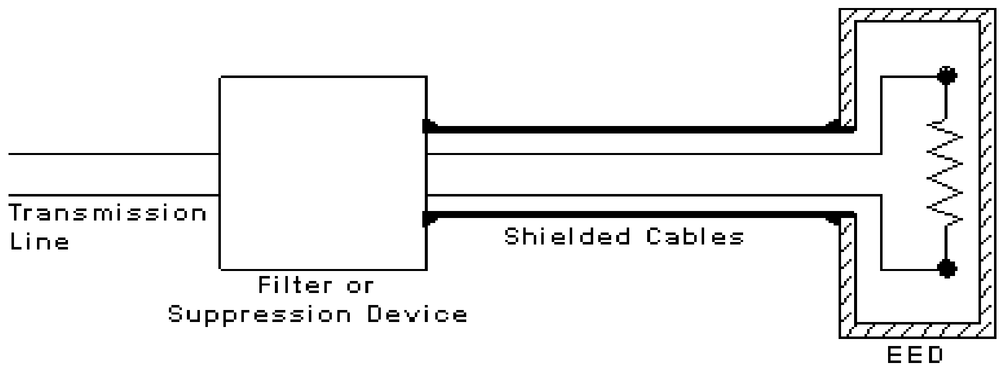


Figure 1-4. Shielded Enclosure with an EMI Filter

1-5. RF ARCING PROTECTION.

The design of circuits associated with systems that have electrical connections exposed to the EME is very important. RF arcs can occur when connectors are mated and unmated, especially for ordnance that may be attached to very large structures or host platforms that are exposed to high-frequency environments. These arcs can generate EM energy throughout the RF spectrum, including low-frequency components that are in the same band as the firing signal, and will even pass through a filter if one is installed. A break in the firing circuit between the arc point and the EID until after the connection is made will circumvent this problem because a direct current path is necessary for an arc to occur. This technique is illustrated in figure 1-5.

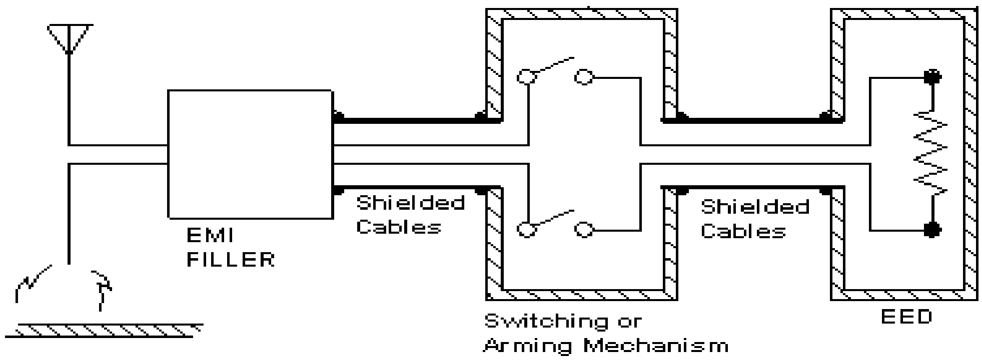


Figure 1-5. Basic Solution to the RF Arcing Problem

1-6. ADDITIONAL DESIGN CONSIDERATIONS.

The goal of the HERO system design is to ensure that the total ordnance system is not susceptible to the EME at any time during its stockpile-to-safe separation sequence. The operational usage of an ordnance system as defined by its stockpile-to-safe separation sequence is divided into six distinct phases: (1) transportation/storage, (2) assembly/ disassembly, (3) handling and loading, (4) staged, (5) platform loaded, and (6) immediate post-launch. Ordnance system immunity to RF susceptibility is achieved through the application of good design practices and quality manufacturing processes. The management of a program to achieve the desired level of HERO performance is described in chapter 9. The test approach to validate the design approach and to determine the degree of susceptibility of the ordnance system to the EME is discussed in chapter 10.

1-7. SUMMARY.

To summarize, there are four basic approaches to solving the HERO problem:

a. Enclose the entire ordnance system in a continuous RF shield.

b. Shield the compartments and the interconnecting cables of the firing circuits.

c. Use an EMI filter in the firing circuit and shield the cable from the filter to the EID.

d. Provide a break in the firing circuit between the filter and the EID for protection from arcs.

One of these approaches, or a suitable combination of them, must be selected early in the design stage and implemented throughout the design, development, and manufacture stages to assure an optimum and economical solution to the HERO problem. Designing an ordnance system that is immune to RF effects is a very challenging and demanding effort. It is the responsibility of the developer to select the approach to be used to determine the attenuation values of the filters and the shielding effectiveness of the enclosure and the cables that will be needed. One way to solve this problem is to consider the ordnance as a receiving system in the EME, and the EID's as the terminating load for this receiver. A good rule of thumb is to provide additional protection so that the total attenuation from the combination of all shielding is 40 dB at 100 kilohertz and increases linearly to 60 dB at 1 megahertz. The attenuation should remain at or above 60 dB from 1 megahertz to 40 gigahertz.

CHAPTER 2

THE ELECTROMAGNETIC HAZARD

2-1. GENERAL.

The electromagnetic susceptibility of ordnance is discussed in this chapter in terms of three major factors: (1) identification and description of the electromagnetic environment (EME) to which the ordnance system may be exposed, (2) possible modes of energy transfer from the EME to the electrically initiated device (EID), and (3) measurement of the EME. Although the information given herein is not essential to the implementation of the principles and guidelines to be established in later chapters, it is presented to give the ordnance system designer an insight into the need and purpose of these principles and guidelines, and to present environmental levels to be used as design goals.

2-2. SUMMARY OF THE ELECTROMAGNETIC ENVIRONMENT.

The available power in the EME created by communications and radar systems at the ordnance system site is a function of the power radiated from the source, the source antenna gain, and the location of the ordnance system relative to the source. The basic relationship of these factors can be derived by reference to an isotropic radiator. An isotropic radiator is a theoretical concept defined as a point source with radiation properties that are uniform in all directions. For an isotropic radiator in free space radiating a power (P_t) in watts, the power density or power per unit area on the surface of a sphere, concentric with the point source and of radius (r) meters, is the total radiated power divided by the area of that sphere, or

$$PD = \frac{P_t}{4\pi r^2} \tag{2-1}$$

where

PD = power density $(watts/meter^2)$.

From this equation, it can be seen that power density in free space decreases inversely as the square of the distance from the radiating source. (This assumes far-field, plane wave conditions.)

If the power source is not an isotropic radiator but radiates with a gain in a given direction, the power density at a point of distance (r) meters in the direction of the gain is

$$PD = \frac{P_t G_t}{4\pi r^2} \quad (2\text{-}2)$$

where

G_t = directional gain of the transmitting antenna (a unitless ratio).

In the far field, the power density and the electric field strength at any point are related by

$$PD = \frac{E^2}{120\pi} \quad (2\text{-}3)$$

or

$$E = \sqrt{120\pi PD} \cong 19.4\sqrt{PD}, \quad (2\text{-}4)$$

where
E = electric field strength $(volts/meter)$.

The factor 120π is known as the intrinsic impedance of free space and is approximately 377 ohms.

If the power density is in $(milliwatts/cm^2)$ and the electrical field strength is desired in volts per meter, the conversion factor of $1 \ \ watt/meter^2 = 0.1 \ (milliwatts/cm^2)$ is used and the equation becomes

$$E = 61.4\sqrt{PD} \quad (2\text{-}5)$$

where
E = volts/meter, and
PD = Power density $(milliwatts/cm^2)$.

The electric field from a transmitter in free space can be computed for any point if the distance to the point, gain of the antenna in the direction of the point, and the power being transmitted are known. Consider the field from a half-wave dipole in free space at a point of distance (r) meters from the antenna in the direction of maximum gain.

From the equations

$$PD = E^2/120\pi \text{ and } \frac{P_t G_t}{4\pi r^2} \quad (2\text{-}6)$$

we have

$$E = \frac{\sqrt{49.2P_t}}{r} = \frac{7.01\sqrt{P_t}}{r} \tag{2-7}$$

where

G_t = 1.64 for a dipole.

The antenna gain is sometimes expressed in decibels (dB). From the definition of dB (dB = 10 log (ratio of two amounts of power)), we have

$$g_t = 10 \log G_t \tag{2-8}$$

or

$$G_t = 10^{gt/10} \tag{2-9}$$

where

g_t = antenna gain in dB.

2-3. SUMMARY OF ENVIRONMENTAL LEVELS.

The degree of electromagnetic (EM) susceptibility of existing ordnance, as determined by analysis of data obtained on HERO tests, is indicated by the maximum safe field curves presented in figures 2-1 through 2-4.

The maximum EME curves in figures 2-1 and 2-2 are based on theoretical and empirical consideration of the receiving characteristics of bare EID's exposed in an EME. These curves represent the worst-case condition which can exist for Naval ordnance (during assembly, disassembly, or testing of EID's). This ordnance is considered by the Navy to be HERO UNSAFE ORDNANCE. The data will be useful in determining the maximum safe fields for bare EID's with lead wires arranged in optimum receiving orientation. There has been no known case of an EID initiating accidentally when the EME level was below the values given by the curves.

The maximum safe EME curves of figures 2-3 and 2-4 represent the safe field strength and power densities for HERO SUSCEPTIBLE ORDNANCE undergoing normal handling and loading operations. These curves are based on experimental results of HERO tests. The boundaries were established by the most susceptible ordnance items (those in which little or no design consideration was given to HERO problems).

Table 2-1 gives the maximum EME that ordnance will encounter during its stockpile-to-safe separation sequence. The trend in both radar and communications equipment toward greater effective radiated power will increase these fields. Past experience can yield some indication of the magnitude of the increase to be expected in the future. For example, early magnetron tubes could supply 10 kW of peak power to a matched antenna. Within a decade, the peak power of magnetron tubes have increased to greater than megawatts.

2-4. POWER LEVELS.

In discussing the power levels of the environment, the effects of modulation on the power level must be considered. Many communication systems use forms of amplitude modulation. If the transmitter is amplitude-modulated, the peak envelope power (PEP) may be as high as four times the PEP of the unmodulated wave. However, this has been taken into account in determining the maximum environmental levels of table 2-1.

Most radar systems use pulse modulation as opposed to the continuous carrier or doppler system. The important parameters of the pulsed system are: pw = pulse width (microseconds), PRF = pulse repetition frequency (hertz), P_p = peak power (watts), and P_a = average power (watts). There is a definite ratio between the peak and the average power that depends on the pulse width and the PRF. This relationship is called the duty cycle and is expressed as follows:

$$\text{Duty cycle} = \frac{\text{average power}}{\text{peak power}} = \frac{P_a}{P_p} \qquad (2\text{-}10)$$

and

Duty cycle = pulse width x pulse repetition frequency = pw x PRF. (2-11)

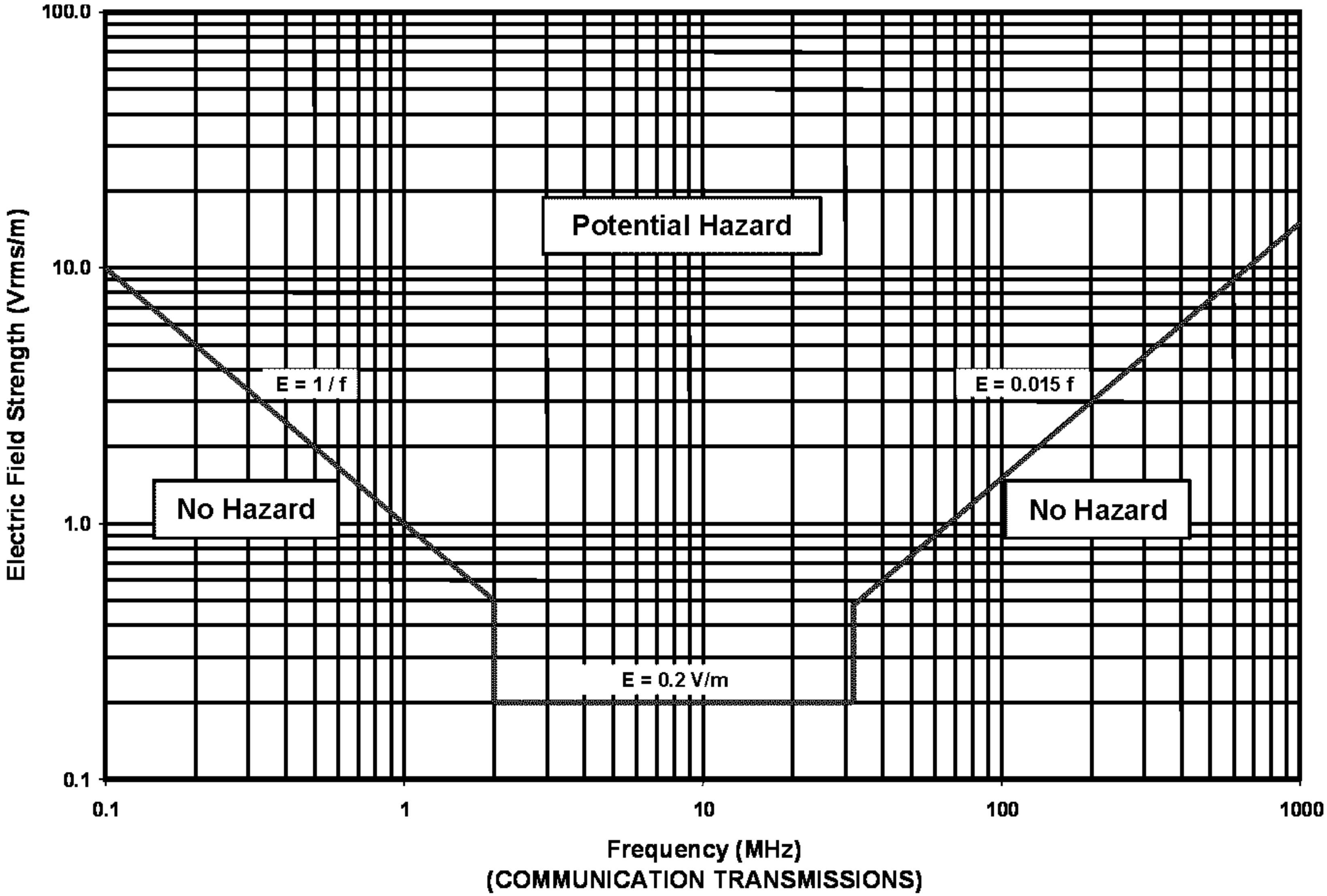


Figure 2-1. Field Intensity Potentially Hazardous to EID's in Optimum Coupling Configurations-Communication Transmissions

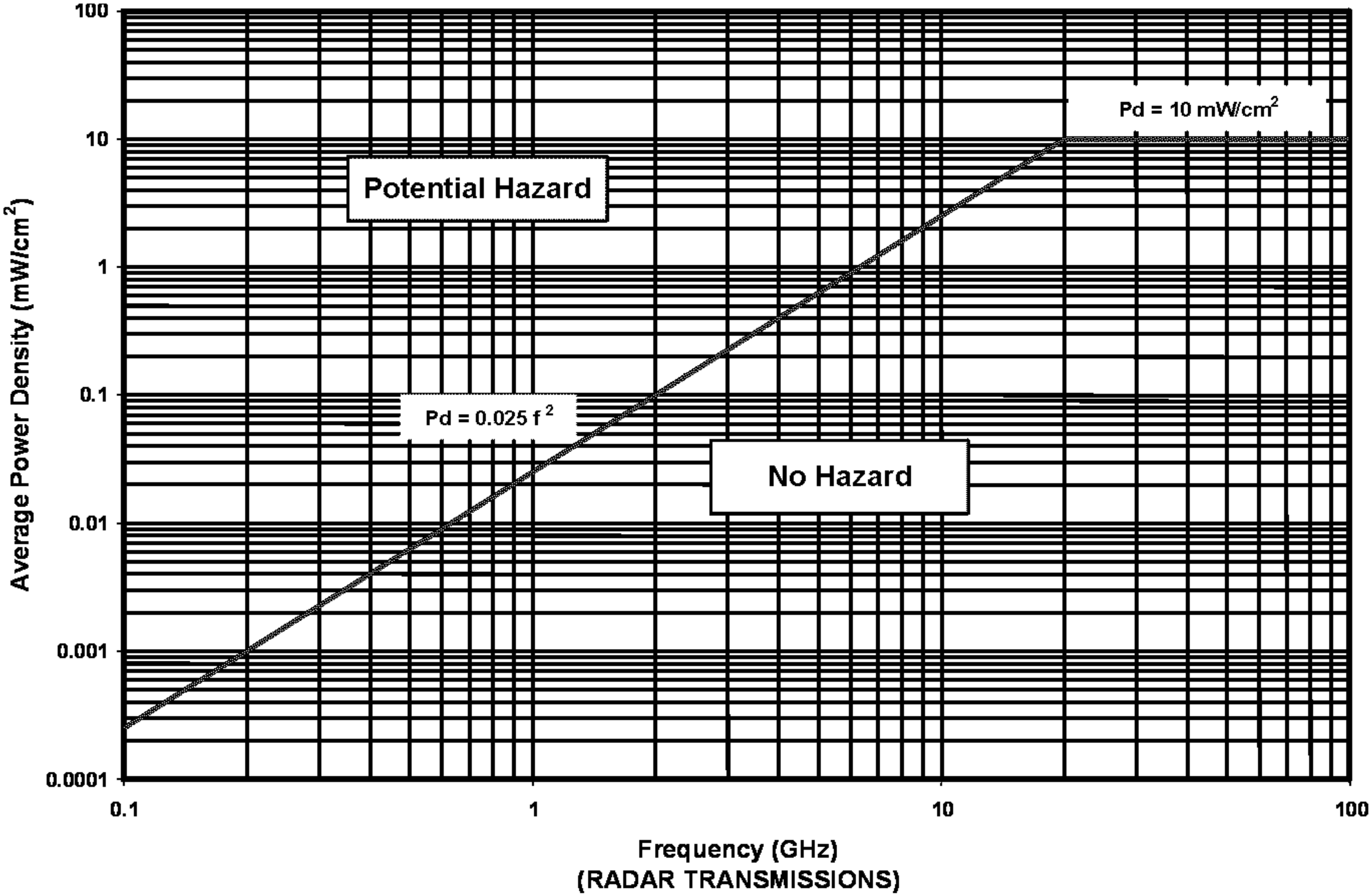


Figure 2-2. Power Density Potentially Hazardous to EID's in Optimum Coupling Configurations-Radar Transmissions

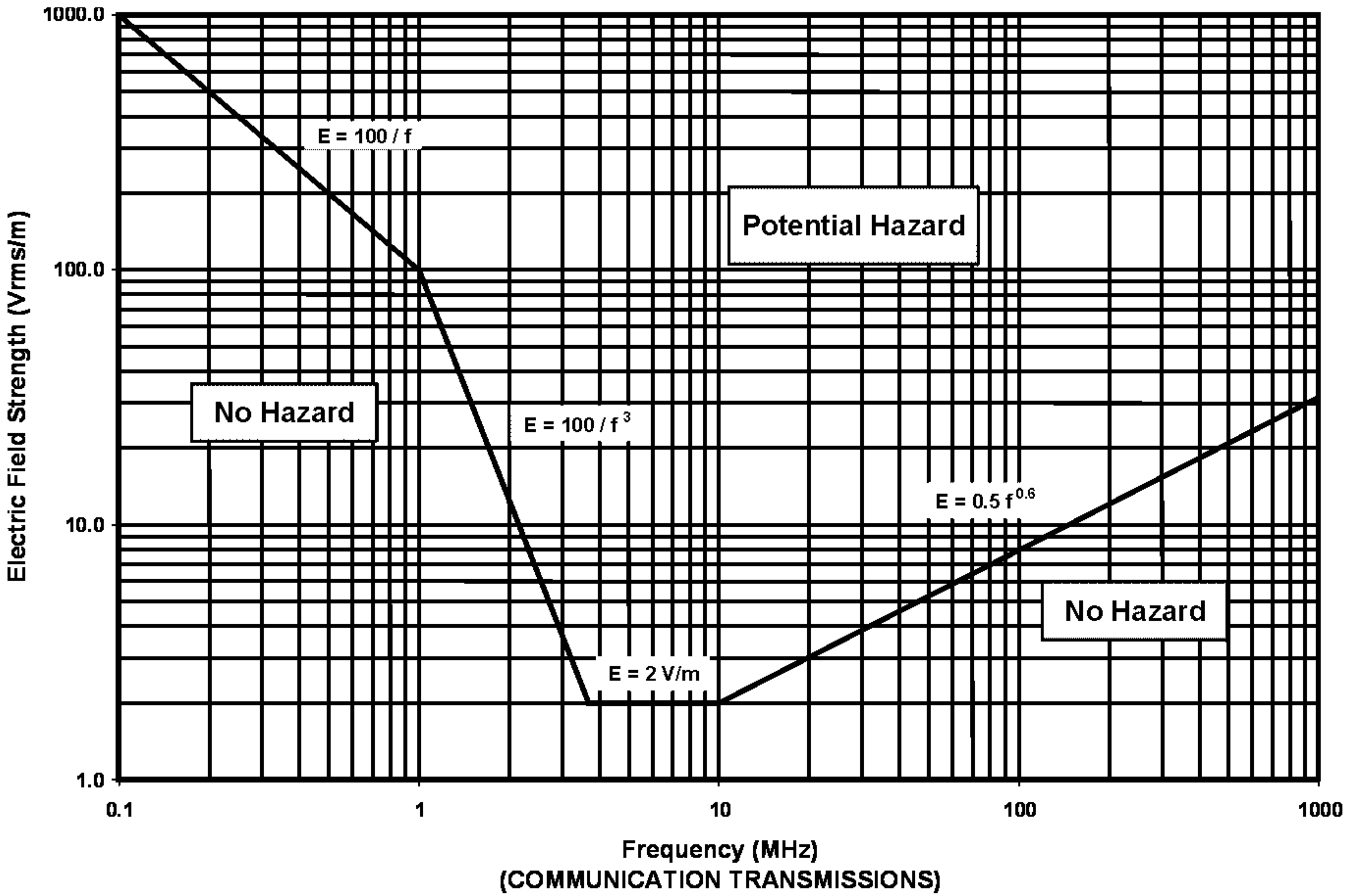


Figure 2-3. Field Intensity Potentially Hazardous to Susceptible Ordnance which Require Special Restrictions-Communication Transmissions

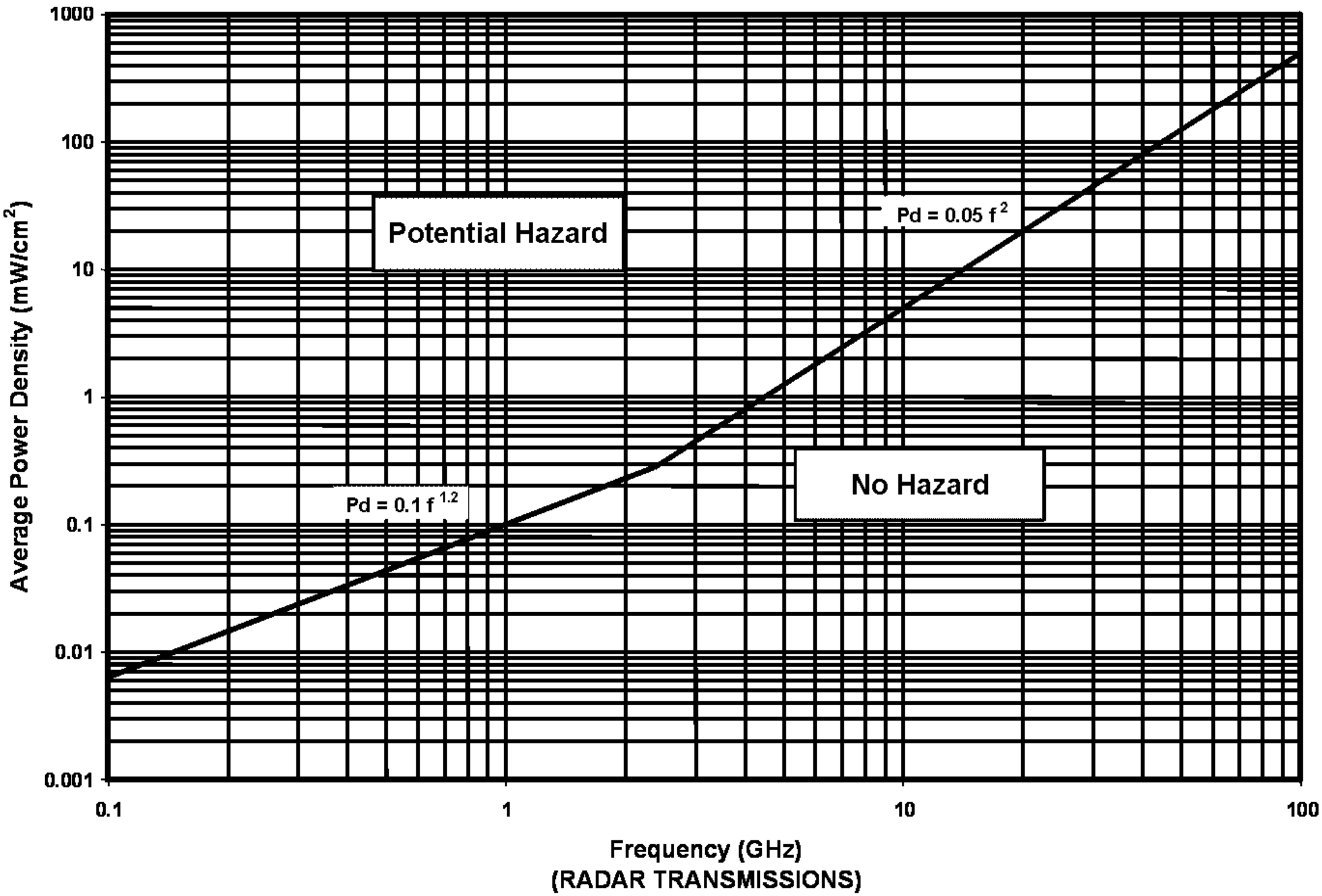


Figure 2-4. Power Density Potentially Hazardous to Susceptible Ordnance which Require Special Restrictions-Radar Transmissions

(These parameters are shown in figure 2-5.) Also, the average power can be computed from the duty ratio as

$$P_a = P_p \; x \text{ duty cycle} = P_p \; x \; pw \; x \; PRF \tag{2-12}$$

or

$$P_a = P_p \, x \, \frac{pw}{T}, \tag{2-13}$$

where

T = time interval of pulses $= 1/PRF$.

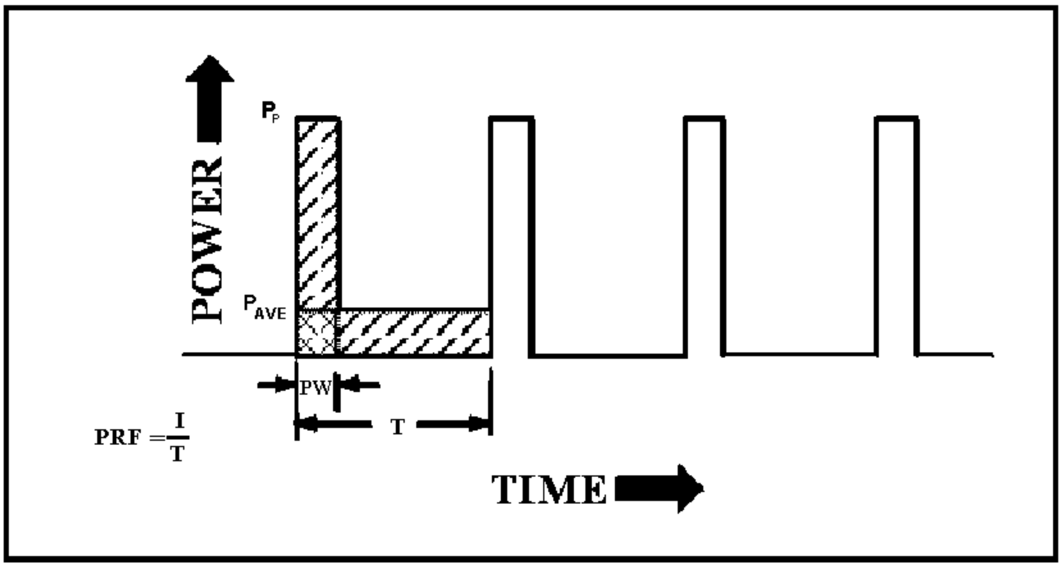


Figure 2-5. RF Pulse Train

2-5. ANTENNAS.

Antennas may be conveniently grouped into two general classes according to the value of the ratio of the antenna's physical size to the wavelength of the transmitted frequency. When this ratio is much greater than unity, the antenna is classed as a large radiator; when it is in the order of unity, the antenna is classed as a small radiator. One type of small radiator is the half-wave dipole. Some of the characteristics of this antenna are shown in figure 2-6. Monopole and long-wire antennas are considered variations of this type.

Table 2-1. Electromagnetic Environment Levels

FREQUENCY (MHz)	FIELD INTENSITY* (Volts (RMS)/Meter)	
	PEAK	AVERAGE
0.1 - 0.6	200	200
0.6 - 2.0	200	200
2.0 - 30	200	200
30 - 150	200	200
150 - 225	3120	270
225 - 400	2830	240
400 - 700	4000	750
700 - 790	3500	240
790 - 1000	3500	610
1000 - 2000	5670	1000
2000 - 2700	21270	850
2700 - 3600	27460	1230
3600 - 4000	21270	850
4000 - 5400	15000	610
5400 - 5900	15000	1230
5900 - 6000	15000	610
6000 - 7900	12650	670
7900 - 8000	12650	810
8000 - 14000	21270	1270
14000 - 18000	21270	614
18000 - 40000	5000	750

*These intensities apply to the smaller of the following field components:
1. The vertical component of the electric field (E).
2. The directional maximum component of the horizontal magnetic field in ampere turns/meter (H), multiplied by 377 ohms.

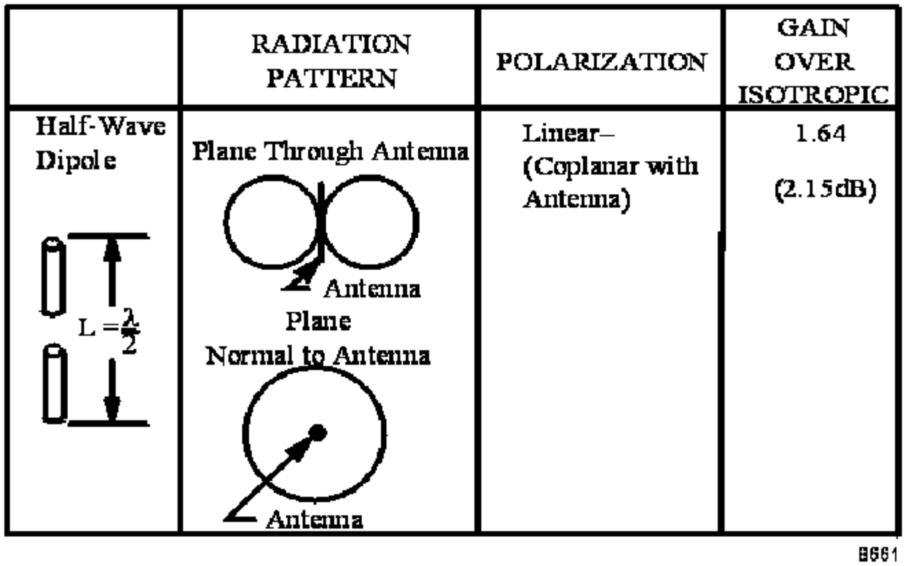


Figure 2-6. Characteristics of a Half-Wave Dipole

Large radiators that are used aboard ship are almost always radar antennas and are most frequently (where high power is concerned) employed with search, height, or guidance radars. Such antennas usually consist of dish-type reflectors. The reflector is designed to alter the phase and amplitude relationships of the feed antenna to focus the radiation at some point in space. Figure 2-7 shows a feed and reflector system typical of those used in radars, together with the associated radiation pattern.

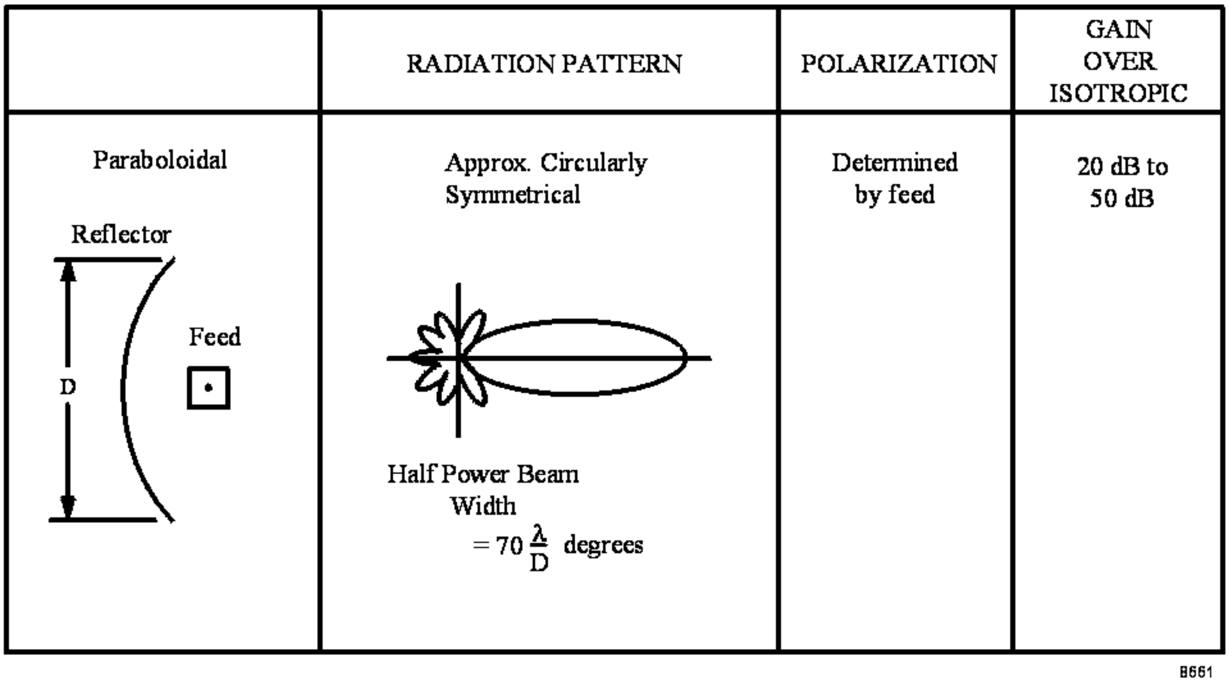
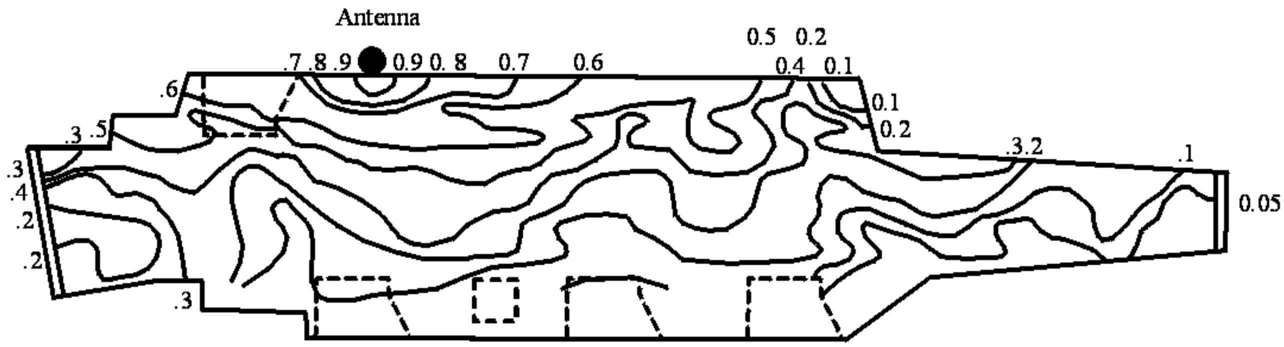


Figure 2-7. Characteristics of a Reflector Antenna

The EME produced by shipboard antennas is important to the HERO problem because a knowledge of the field strength is necessary for determining the amount and type of protection needed for the ordnance. (See figures 2-1 through 2-4.) Unfortunately, only in the region where the antenna's field appears as a plane wave, decreasing as an inverse function of the distance from the antenna $(E = f(1/r))$, can any positive measurements be made or field intensity relationship be established. This region is known as the far field or Fraunhofer region. Knowledge of the intensity at one point in this space can lead to an accurate extrapolation of the intensity at another point. It is in this region of an antenna's field that there is also a definite relationship between the electric and magnetic fields. They are related by the equation

$$E = 120\pi H.$$

Even though some areas of a ship are in the far field of an antenna, additional complications are introduced by reflections and discontinuities in the propagating medium. Figure 2-8 depicts typical field strength contours on the deck of an aircraft carrier. The irregularity of the shape of the contours suggests the difficulty of predicting an EME. The contours shown are a measure of the electric field which was generated by a single transmitter feeding one monopole antenna located at the edge of the carrier deck. The change in the pattern that would occur with the addition of another transmitting system is virtually unpredictable.



Vertical component of the E field 3 feet above the deck expressed in volts per meter for 1 watt radiated power at a given communication frequency.

Figure 2-8. Typical Field Strength Contours on a Carrier Deck

The place where the far field of an antenna begins is not exactly defined. It is an arbitrarily chosen region where the previously described effects begin to be evident. For small radiators, it is usually considered to begin at a distance of approximately two wavelengths from the antenna. For large radiators, $2D^2/\lambda$ (where D is the largest dimension of the antenna) is commonly accepted.

The near field of any antenna is the region or space between the antenna and the beginning of the far field. It is composed of the combination of effects from two regions, the inductive region and the Fresnel region. The inductive region is considered to be significant up to one wavelength from all antennas. The Fresnel region (or interference region) is considered to begin one wavelength from the antenna, and its noticeable effects extend to the beginning of the far field. The near field of shipboard communication transmitters utilizing electrically small radiators is such that there is little effect due to Fresnel interference; consequently, the near field is considered to be made up entirely of the induction field.

For shipboard radar antennas, the Fresnel interference is significant and cannot be ignored. Since the induction field is only significant to distances comparable to a wavelength, this usually amounts to no more than a few centimeters for radar antennas. Insofar as HERO is concerned, the only applicable consideration for radar fields is the Fresnel region, since it is unlikely that ordnance will be employed at the aperture of shipboard radar antennas.

2-6. ELECTROMAGNETIC ENERGY TRANSFER.

The power received by an antenna in a uniform field is a function of its effective area and the power density at the antenna location. That is,

$$P_r = A_{er}PD, \qquad (2\text{-}14)$$

where

P_r is the power $(watts)$ delivered to the load impedance across the antenna terminals, and A_{er} is the effective area of the antenna $(meters^2)$.

The effective area of a receiving antenna is given as

$$A_{er} = \frac{G_r\lambda^2}{4\pi}, \tag{2-15}$$

where

G_r = gain of receiving antenna, and

λ = wavelength in meters = 300/frequency in megahertz.

This expression is for the maximum effective area of an antenna and it occurs only when the antenna is matched to its load. Therefore,

$$P_r = \frac{G_r\lambda^2 PD}{4\pi} \tag{2-16}$$

if the power density at the receiving antenna is known, or

$$P_r = \frac{G_r G_t P_t \lambda^2}{(4\pi r)^2} \tag{2-17}$$

if the power transmitted, distance to transmitting antenna, and gain of transmitting antenna are known.

These equations are valid only when the load is matched to the impedance of the antenna, since the expression for A_{er} is for maximum effective area, and occurs only when the load and antenna are matched.

The following sample calculation illustrates the principle for determining the induced current in an EED bridgewire that terminates a half-wave resonant dipole antenna. We must assume the following conditions:

a. The lead wire length (AB and CD are arranged so that a half-wave dipole is formed (see figure 2-9). This antenna is terminated in a 1-ohm EED.

b. The characteristic impedance and length of the transmission line (formed by BE and CF) are such that the 1-ohm load is matched to the antenna. The losses of the transmission line are neglected.

c. The antenna gain G_r relative to an isotropic antenna is taken as 1.64 (see figure 2-6).

d. The field strength is assumed to be 100 volts per meter at 30 megahertz.

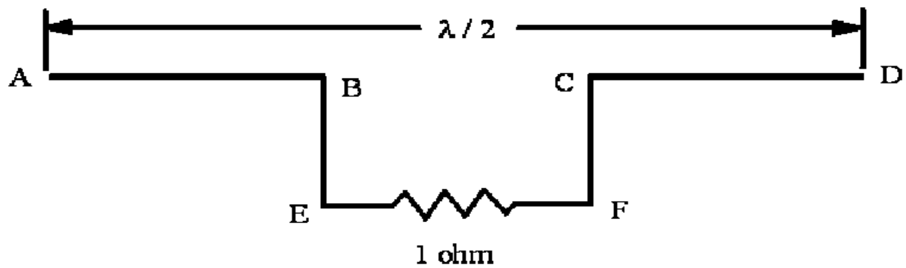


Figure 2-9. An EED Matched to a Dipole Antenna

Use the equation

$$P_r = \frac{G_r \lambda^2 PD}{4\pi} \qquad (2\text{-}18)$$

where

$PD = E^2/120\pi = 26.5\ watts/meter^2$ (this assumes far-field conditions),

$G_r = 1.64$ (gain ration for a half-wave dipole antenna with a 2.1 dB gain relative to an isotropic radiator), and

$\lambda = 10\ meters\ [\lambda = 300/frequency(MHz)]$

Therefore,

$$P_r = \frac{1.64x(10)^2 x26.5}{4\pi}\text{, and} \qquad (2\text{-}19)$$

$P_r = 345.8\ watts$.

The current in the bridgewire of the EED is calculated from

$$P = I^2 R$$

where

P = power $(watts)$,

R = resistance $(ohms)$, and

I = current $(amperes)$.

Thus, for $R = 1\ ohm$ (a typical value for EED's), we have:

$$I = (P/R)^{0.5} = (345.8/1.0)^{0.5},$$

$$I = 18.6\ amperes.$$

The induced current in an EED bridgewire as previously calculated represents a worst-case situation where all protection normally found in the ordnance, such as shielded cables and shielded enclosures, have been omitted. Also, all losses due to transmission line and impedance mismatches have been ignored. It is a theoretical method for obtaining maximum values. The current in the bridgewire has never been found to exceed the value calculated by this method.

The structural enclosure of an ordnance item provides some EM shielding for the enclosed EEDs. In actual conditions found in ordnance, the problem of analyzing the details of the complete mechanism of the transfer of energy from the EME to an EED does not lend itself to a straightforward theoretical solution. However, it is unlikely that the worst-case example could occur in the completely assembled ordnance system.

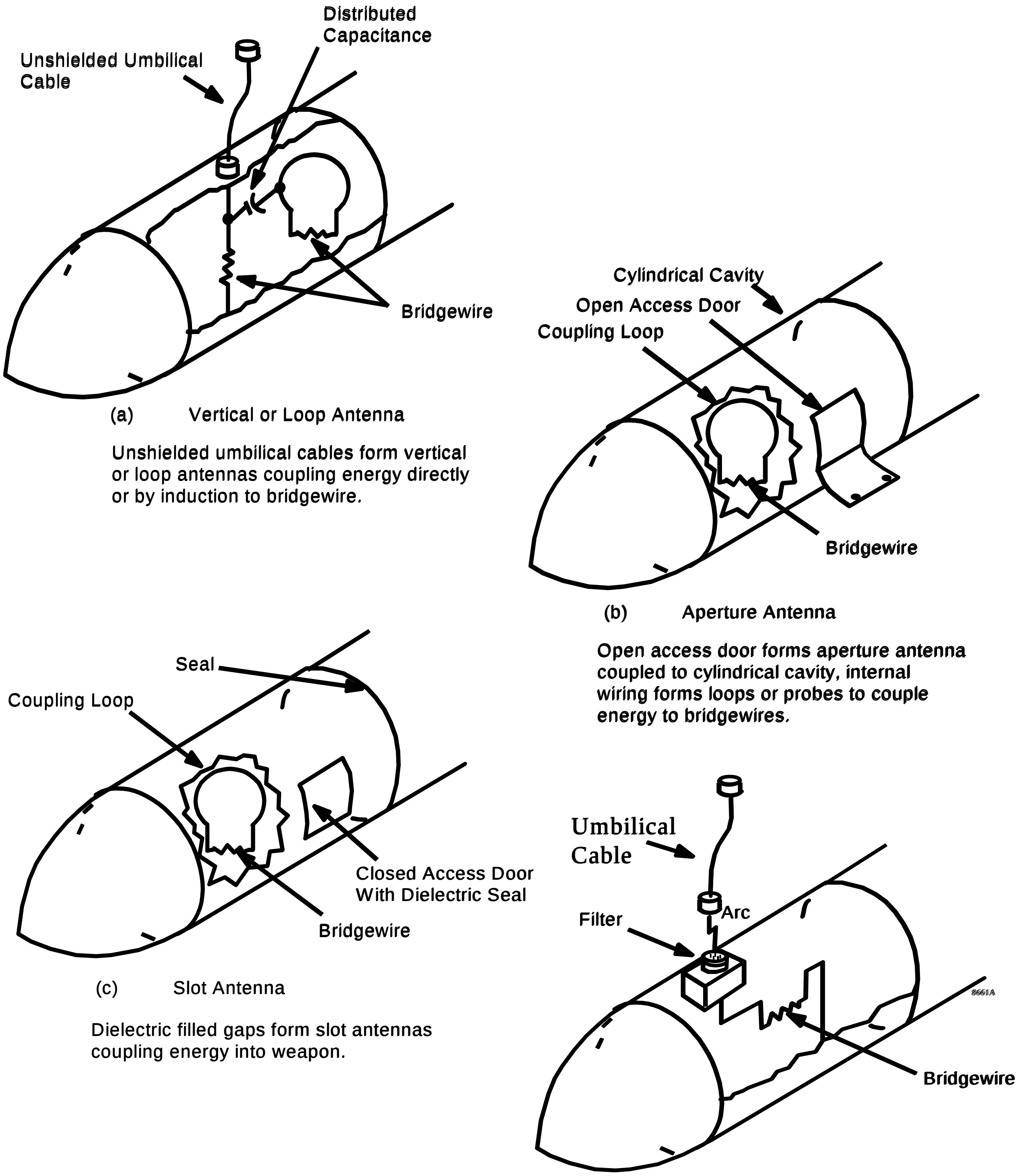
The exterior of the ordnance may be energized either by incident fields from external sources or by direct coupling from its own internal sources. Whatever the source, the surface distribution of current and charge may exhibit stationary patterns depending on the method of excitation, the wavelength of the excitation current, and the geometry of the ordnance. These patterns are usually very complicated.

In electrical and mechanical form, the receiving antennas of the ordnance system that contribute to the HERO problem are not necessarily recognizable as antennas. They may be the aircraft, launchers, umbilical cables, access doors and hatches, or discontinuities in shields, but they, nevertheless, function as linear antennas, current loops, or cavity and slot-aperture antennas.

Some of the ways in which umbilical cables, apertures, and discontinuities in the shield can function as receiving antennas for EM energy are shown in figure 2-10. Panel (a) illustrates an umbilical cable as the receiving antenna (vertical or loop) and an internal loop antenna consisting of an EED and its associated wiring. External cables can act as effective receiving antennas when exposed to an EME permitting the transfer of RF current into the ordnance which can couple directly or inductively into an EED bridgewire. This type of receiving antenna can be an effective receiver at communication frequencies, depending on the length of the external cables and their connections.

Panels (b) and (c) of figure 2-10 illustrate apertures in the ordnance skin acting as receiving antennas. These apertures are effective receiving antennas at frequencies where their dimensions approach one wavelength, and the amount of energy transferred from the environment into the cavity becomes more pronounced. This occurs most often at radar frequencies. The energy is coupled from the fields developed in the cavity to the bridgewire by capacitive and inductive means.

Panel (d) illustrates energy transfer occurring as a result of an arc. When a connection is either made or broken between any two ordnance elements having different electrical potentials (e.g., connectors between ordnance and launcher or between ordnance and test equipment), arcs occur which can produce large amounts of energy at all frequencies including dc and low frequency ranges. If arcs occur in the firing circuits, this energy can be delivered to an EED even if the EED is protected by an electromagnetic interference (EMI) filter (see chapter 8).



(a) Vertical or Loop Antenna

Unshielded umbilical cables form vertical or loop antennas coupling energy directly or by induction to bridgewire.

(b) Aperture Antenna

Open access door forms aperture antenna coupled to cylindrical cavity, internal wiring forms loops or probes to couple energy to bridgewires.

(c) Slot Antenna

Dielectric filled gaps form slot antennas coupling energy into weapon.

(d) Conduction of RF Arc

The RF arc striking an unprotected EID lead can cause low frequency and DC currents to flow in the EID circuit even though the filter is used for protection.

Figure 2-10. Ways in which Ordnance Components can Function as Receiving Antennas

Under any of the conditions illustrated in figure 2-10, the energy transfer can be increased by the presence of personnel in proximity to the ordnance. The human body displays receiving antenna characteristics and can thus increase the efficiency of the transfer path of EM energy to the susceptible portions of the ordnance.

Attempts to analyze the amount of energy coupling by a theoretical study of apertures, lead-to-lead intercoupling, lengths of wires, impedance match or mismatch, and effectiveness of shielding, have all failed, due to the complexity of the problem.

2-7. EME MEASUREMENTS.

The parameters used to describe the EME are generally the following:

E = electric field $(volts/meter)$,

H = magnetic field $(ampere\ turns/meter)$, and

PD = power density $(watts/meter^2)$.

The polarization of a radiating source is defined in terms of the orientation of the electric field with respect to a reference plane (usually the surface of the Earth). Accordingly, the polarization is not restricted to linear polarization in the horizontal and vertical planes of propagation, but can contain both horizontal and vertical components, establishing elliptical polarization.

The magnetic field is not restricted to any one plane of propagation, but follows the polarization forms of the electric field.

The instantaneous power flow per unit area from a radiating source may be represented by the Polynting vector:

$$P_i = \bar{E} x \bar{H}. \qquad (2\text{-}20)$$

To determine the outward power flow from a given radiator to a point in space, the preceding expression must be integrated over a complete cycle. The form most useful in calculating the power density of a source of radiation is the complex Polynting vector

$$PD = 1/2 Re(\bar{E} x \bar{H}^*), \qquad (2\text{-}21)$$

where $\bar{H}^*$ is the complex conjugate of $\bar{H}$.

In the far field of a radiation source, $\bar{E}$ and $\bar{H}$ are transverse to the direction of propagation, and are complicated only by the nature of their polarization, which may consist of both vertical and horizontal components. In the near field, $\bar{E}$ and $\bar{H}$ are further complicated by having components that are not transverse to the direction of propagation and by the existence of the reactive fields of the radiation source.

Field measuring devices can be divided into three basic categories: (1) those sensitive to the electric component, (2) those sensitive to the magnetic component, and (3) those sensitive to the power density. An ideal measuring device would be sensitive to the power density and capable of summing the contribution of all field components at the point of measurement. Since this ideal is not easily realized, the measuring techniques must compensate for the limitations in measurement devices.

At communications frequencies, it is common to use field-measuring equipment that indicates either the electric field intensity $(E)(volts/meter)$ or the magnetic field strength (H) $(ampere\ turns/meter)$. At radar frequencies, it is common to use equipment that measures power density. Although the above is not an absolute rule, the types of detectors and their frequency characteristics have made it convenient.

Some measuring devices employ electric or magnetic field detectors and electronically convert the indication to power density or to the unmeasured electric or magnetic component of the field. It should be noted that the conversion of $\bar{E}$ or $\bar{H}$ directly to PD is valid only when the relationship between E and H is known. The relationship of E/H in the near field is very complex and difficult to measure.

CHAPTER 3

ELECTRICALLY INITIATED DEVICES

3-1. GENERAL.

The family of components generally referred to as electrically initiated devices (EID's) is used to convert an electrical input to a desired energetic output. This family of components includes (but may not be limited to) devices that use bridge-type transducers comprised of wire, carbon, conductive composition, semiconductor, or foil, and non-bridge type devices such as laser initiators.

EIDs generally fall within two broad categories based upon their output response to an electrical input; devices that produce a dynamic mechanical or thermal response, and devices that produce an explosive response. Devices that produce a dynamic mechanical or thermal response include burnwires, fusible links, and heat-activated release mechanisms. Those that produce an explosive response are known as electroexplosive devices (EEDs). These types include ignitors, detonators, squibs, explosive actuators, and explosive switches.

The energy source used to initiate these types of devices is normally an ac or dc firing circuit. However, any electrical energy, including electromagnetic (EM) energy coupled to the device or its firing circuit, can initiate it. This is the basic HERO problem. This chapter describes the manner in which EID's function and discusses the susceptibility of such devices to EM energy. The advantages and disadvantages of types of available EID's are also provided. The purpose is to give the designer background information essential to understanding the HERO problem as it relates to EID's and to assist the designer in selecting an EID that is suited both to the requirements of the weapon system and its protection from HERO.

3-2. EFFECTS OF ELECTROMAGNETIC ENERGY COUPLING.

3-2.1. INADVERTENT INITIATION. There are two primary modes of EID excitation associated with the coupling of EM energy into an ordnance system. One of the modes stems from EM energy coupling directly to the EID itself, causing it to function unintentionally. Secondly, EM energy can be coupled into and activate the EID firing circuit causing the circuit to function as designed; i.e., fire the EID. Actuation of an EID either directly or indirectly causing the ordnance system to operate prematurely can create a hazardous situation (safety concern) or performance degradation. If the EID initiates either out of sequence or before the system is armed, the effectiveness of the system may be reduced, thereby creating a reliability concern.

3-2.2. DUDDING. The application or repeated application of EM energy below that required for initiation of an EID can cause desensitization of the energetic material in the area around the bridge. In fact, the energetic material can become so desensitized that the device will not initiate when the appropriate electrical stimuli are applied to the device, resulting in a dud EID.

Metallic and semiconductor bridges can be burned through or damaged, also rendering the EID useless. This dudding effect is undesirable and creates a reliability concern.

3-2.3. THERMAL STACKING. In pulsed electromagnetic environments (EME's) such as the one that results from radar transmission, there occurs a phenomenon called "thermal stacking," which can increase the likelihood of inadvertent initiation or dudding. This phenomenon is limited to slower responding EID's such as those using wire bridges. The heat generated by a single pulse of energy may be insufficient to initiate the EID, but if the time between pulses is shorter than the thermal time constant of the device, successive pulses can progressively elevate the bridge temperature until the initiation temperature is reached. Figure 3-1, in which the temperature increases are shown graphically, demonstrates that the temperature will rise from the ambient level until it reaches a final equilibrium point, after which no further increases will occur. This final temperature, which is a function of pulse amplitude, pulse duration, repetition rate (duty ratio), and the thermal time constant, may be sufficiently high to cause dudding or even to initiate the EID. In considering the hazard in pulsed environments, the effects of thermal stacking must be taken into account. The maximum safe power densities indicated by figures 2-2 and 2-4 account for the effect of thermal stacking.

3-3. MODES OF RF EXCITATION.

There are two modes of unwanted radio frequency (RF) excitation in an EID or its firing circuit, the differential mode and the coaxial mode. In the differential mode, the unwanted signal is introduced through the same path as that of the desired firing signal. (That is, the induced current in one wire of the EID flows toward the device, and the current in the other wire flows away from the device.) In coaxial mode excitation, the unwanted signal path is only partly common with the desired firing signal path (that is, the induced current flows toward the EID in both wires); ground loops through the explosive material, capacitive, or inductive coupling usually serve as the return path.

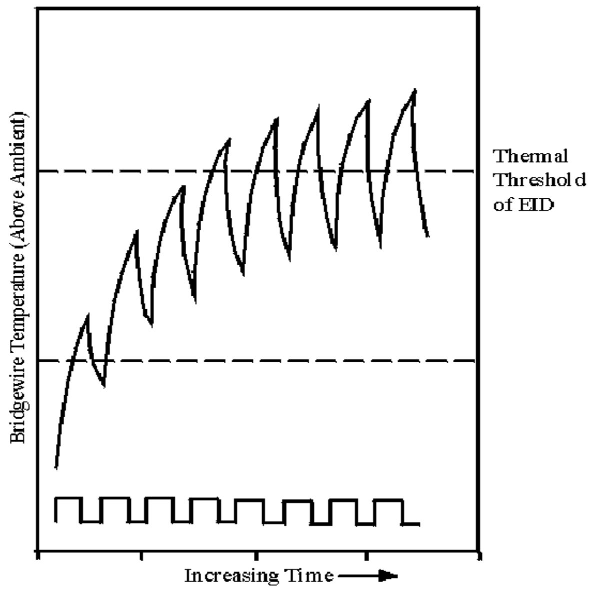


Figure 3-1. Temperature Increases Due to Thermal Stacking

Differential mode RF excitation occurs in two-wire firing circuits. The EM energy propagates to the EID or its firing circuit between two wires in the same manner as the normal ac or dc firing current. This will cause joule (resistance) heating of the bridge material, thereby causing inadvertent initiation or dudding of the EID. Figure 3-2 illustrates the differential mode of excitation. In this mode it might appear that if a large mismatch of impedance occurs between the EID or its firing circuit input and the transmission line, which is usually the case, most of the EM energy would be reflected at the EID or firing circuit input. Although most of the energy is reflected, enough can be transmitted to produce a hazardous condition.

In a coaxial firing system (figure 3-3), the energy propagates to the EID or its firing circuit along one leg or both legs of the normal ac or dc firing paths. The unwanted signal circuit is completed via a ground loop or capacitive/inductive coupling at the EID. This can be visualized easily by considering a wire or metal rod center conductor contained inside a cylindrical conductor, such as a shield, that is concentric with it. And, the EID bridge material is connected at each end to the center and outer conductors. The unwanted signal voltage appears across the EID bridge material as a result of an undesired path that occurs between the center and outer conductors; either through a ground loop or capacitive/inductive coupling. Here again, heat in the bridge material is generated just as the intended ac or dc firing current does.

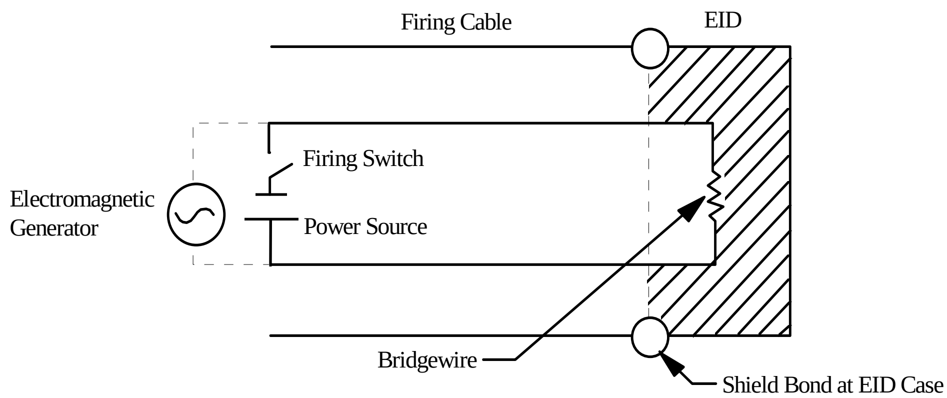


Figure 3-2. Differential Mode of RF Excitation in a Two-Wire Firing System

The coaxial mode of RF excitation can also occur when using a two-wire, balanced shielded system through any high impedance connection in the shield continuity (figure 3-4). In this case, the two lead wires serve as the center conductor and the shield serves as the outer one. In a two-wire balanced system, energy transferred to the EID in the coaxial mode will cause a high potential to be developed from the bridge through the explosive mix, to the EID case. This can cause arcs to occur in the explosive mix or can cause dielectric heating of the mix.

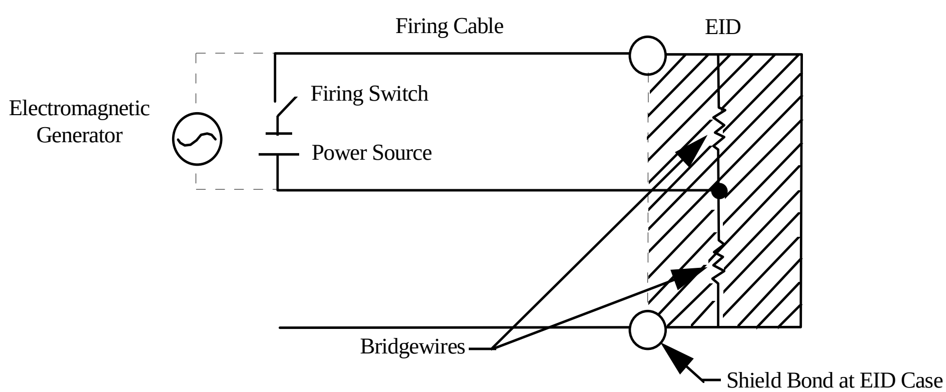


Figure 3-3. Coaxial Mode of RF Excitation in a Coaxial Firing System

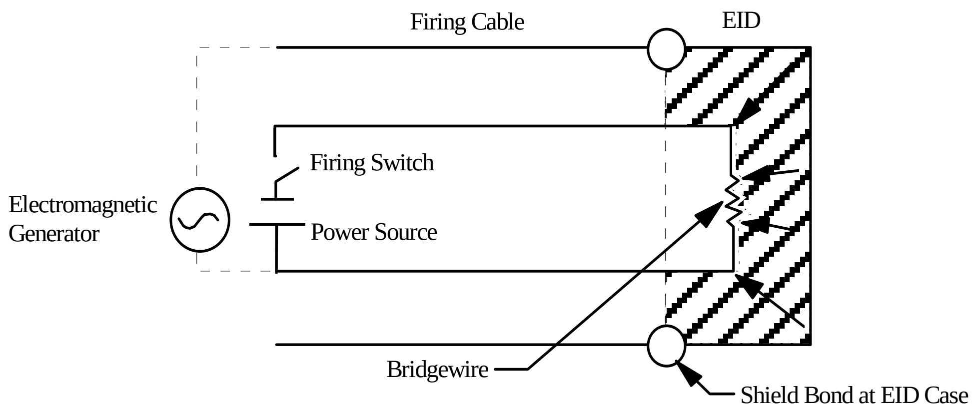


Figure 3-4. Coaxial Mode of RF Excitation in a Two-Wire Firing System

3-4. DESIGN FACTORS AFFECTING EID SELECTION.

In selecting an EID, the designer should be aware of the inherent design features of the various types of EID's as they affect the weapon's susceptibility to EM radiation. Brief descriptions of the types of EID's presently available to the designer with a discussion of methods of determining EID sensitivity are provided in the following paragraphs.

3-5. AVAILABLE TYPES OF EID'S.

3-5.1. HOT BRIDGEWIRE DEVICES. Hot bridgewire devices are used extensively in Navy and joint ordnance for a wide variety of applications. They can take a large number of different configurations, including both NEI's and EED's, but their essential nature remains the same. An ac or dc current is passed through a resistive wire in order to generate joule heating that effects an energetic response.

An example of a hot bridgewire EED, the type most commonly used, is shown in figure 3-5. An EED of this type is normally initiated by heating the bridgewire with an electric current, thus initiating the primary charge surrounding it. In systems where EED's are used to detonate secondary explosives, the primary charge sets off a booster charge, which in turn sets off the main charge. Four common hot bridgewire EED circuits are shown in figure 3-6. Types A and B are generally preferred for protection against HERO, while the use of C and D is generally discouraged.

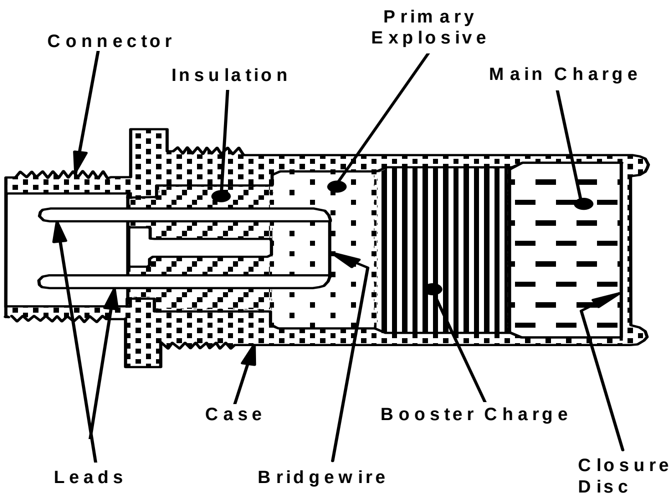


Figure 3-5. Example of Hot Bridgewire EED

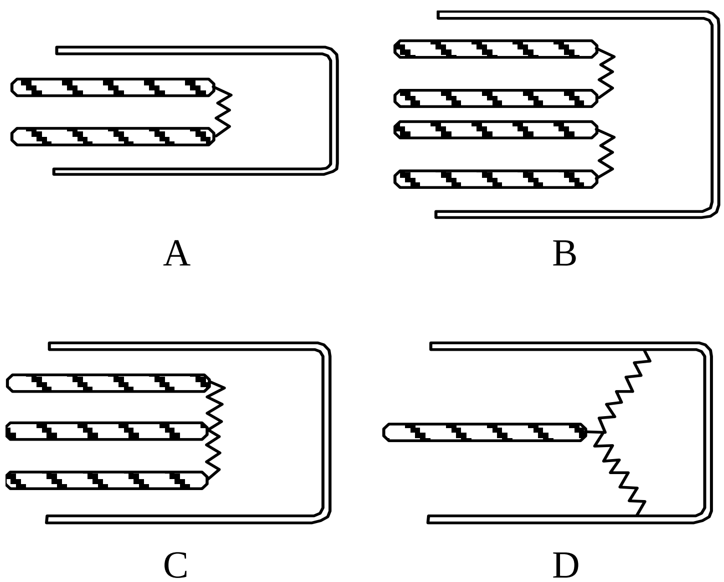


Figure 3-6. Four Types of Hot Bridgewire EED's Circuits

As the name implies, NEI's (see figures 3-7A and 3-7B) do not contain an explosive or pyrotechnic mix. Rather than producing a fast chemical reaction, these devices use a slower "burn resistor." In this configuration, a heated resistor is used to trigger a mechanical process (e.g., burning a string that releases a lever, melting a hole in a pressurized bladder). This process frequently serves to perform the same functions as its explosive counterpart.

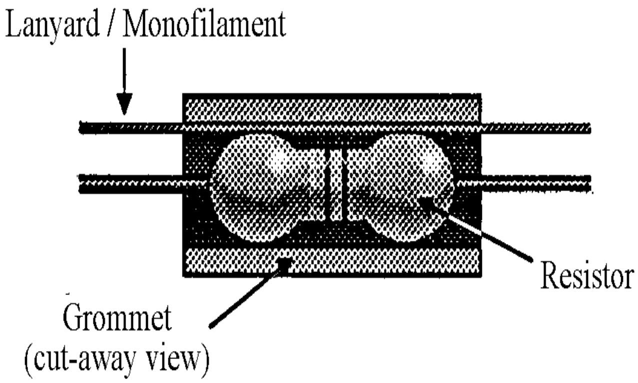


Figure 3-7A. Metal Oxide Burn Resistor

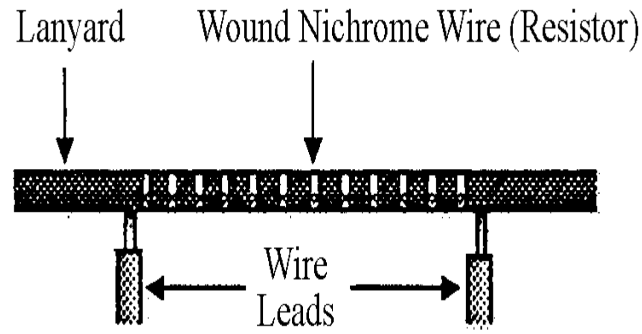


Figure 3-7B. Wound Nichrome Burn Resistor

The 1 ampere/1 watt requirement of MIL-I-23659C for hot bridgewire devices may serve to reduce the hazard from EM energy in proportion to the increase in the power required to fire the EID. However, adherence to this requirement alone will not solve the HERO problem. It is apparent from the maximum safe EME curves (figures 2-3 and 2-4) and the maximum environmental levels (table 2-1) that the potential hazard could not be eliminated for some systems even if 1 ampere/1 watt EID's were used in these systems.

Selection criteria for a 1 ampere/1 watt EID should consider design techniques used by the manufacturer to conform to the no-fire stimuli requirements of MIL-I-23659C. Occasionally, the heat dissipation requirements of such an EID are achieved by introducing metallic materials into the explosive mix or baseplug of the device. The presence of these materials may provide a coaxial-mode current path for EM energy from a firing lead through the explosive mix or baseplug to the case of the EID. The device may be more sensitive to the EM energy through this mode than to the intended (pin-to-pin) firing mode.

3-5.2. EXPLODING BRIDGEWIRE DEVICES. The physical appearance of an exploding bridgewire (EBW) device is similar to that of the more conventional hot bridgewire type. The major difference is the absence of the sensitive primary explosive at the bridge. The operation of the EBW uses thermal and mechanical phenomena that result from the dissipation of a large amount of electrical energy that has been rapidly applied to the bridgewire. Due to the small cross-section of the wire, the current heats the wire material through the melting, boiling, and vaporization phases. The firing stimulus must be applied at such a fast rate that the wire material phase change is restricted due to inertia. When this inertia is overcome, vaporization of the wire occurs as an explosion giving off thermal energy and a shock wave. The exploding bridgewire directly initiates a secondary explosive such as PETN or RDX. This elimination of the primary explosive greatly reduces the sensitivity of the EBW. In order to initiate an EBW, a current in the order of 200 amperes must be applied in microseconds. In general, total energy used is 2 joules; however, EBW detonators can detonate at less than 0.2 joules total energy if the proper circuit parameters are used. It is unlikely that sufficient energy could be accidently transferred from an EME to the device to cause initiation. However, the power transferred by an EME to the device could induce sufficient current to cause the bridgewire to break and dud the

EBW. Under conditions where arcing occurs, it is possible for the secondary explosive next to the bridgewire to deflagrate at relatively low currents.

3-5.3. EXPLODING FOIL INITIATORS. The Exploding Foil Initiator (EFI or slapper detonator) is distinguished from typical EED's by the physical separation of the "foil" and the primary explosive. Operation of the EFI is achieved by applying an electrical pulse sufficient to vaporize the foil. The vapor propels a flyer (usually made of kapton or mylar) along a barrel until it strikes the explosive, causing initiation. The electrical energy required is about 1 joule, delivered in a short pulse, with an associated power of about 5 MW. As a result, the risk of an RF-induced initiation is remote. However, the foil itself is merely a conductor (usually copper or aluminum) and may be susceptible to RF-induced damage, which could affect the reliability of the device. [Since the foil is small, the thermal capacity is small and thus the thermal response for the EFI is fast (on the order of microseconds) indicating a concern over peak power sensitivity.]

3-5.4. CONDUCTIVE MIX INITIATORS. In the conductive mix initiator, the firing current is carried by the explosive mix rather than by a bridgewire. The current path is a powdered conductive material, usually graphite, mixed with primary explosive. Electrical current is passed between the leads through the conductive-explosive mix. The flow of current causes "hot spot" heating that brings the explosive mix to its initiation temperature.

The voltage required for firing a conductive mix initiator varies from 10 to 50 volts. Firing times (3 to 10 microseconds) are much shorter than for HBW devices because the thermal time constant is much smaller. The energy requirements are small (as low as 10 ergs).

A number of design problems are associated with the conductive mix initiator. These problems are manifested in the quality control and production of the initiator rather than in theoretical concepts of design. The resistance of the mix varies widely with the density, homogeneity, and consolidation pressure of the mixture of explosive and conductive particles. The resistance is usually widely variable, in the order of hundreds of ohms, and seems to be particularly well-matched to the induced RF currents. Therefore, conductive mix initiators are not recommended for use.

3-5.5. CARBON BRIDGE EID's. The carbon bridge initiator has been used in military systems since 1950. In the carbon bridge EID, the metal bridgewire is replaced by a conducting bridge of carbon. Colloidal graphite serves as a bridge between two closely spaced electrodes. The primary advantages of the carbon bridge over typical wire bridge EID's are its sensitivity and reliability, which are advantageous in fuze applications where the initiation stimulus is sometimes supplied by low-power piezoelectric crystals. The construction of the carbon bridge detonator is nearly identical to the bridgewire devices, with the exception that the bridge is a deposited carbon-film element vice a conventional wire. The acceptable bridge resistance can range from 1000 to 10,000 ohms in a typical carbon bridge EID (e.g., T62 detonator), indicating a wide statistical variation in electrical characteristics among devices. These devices are also extremely sensitive to electrical stimuli, with maximum no-fire energies (MNFE's) as low as 0.11 μJ for the T62 detonator (i.e., capacitor discharge sensitivity at 0.05 percent probability of firing with 90 percent confidence is 10 V with a 0.0022 μF capacitor). In addition, an estimate of the thermal time constant for these devices (taken from firing voltage versus pulse duration curves) is about 200 μs, indicating the need for some consideration of *peak* power effects in radiated environments (e.g., duty cycles and transients).

Because of the extreme electrical sensitivity and the resulting RF susceptibility of these devices, some initiator specifications preclude or discourage the use of carbon bridge devices:

a. MIL-HDBK-1512 (USAF) - "5.14.2 Carbon Bridgewires. Electroexplosive initiators using carbon bridgewires are prohibited."

b. MIL-I-23659C - "3.3.7 Bridge Material. Carbon shall not be used as bridge material unless specifically approved by the cognizant Government design Agency."

In spite of these recommendations, carbon bridge devices are still being used in a number of Navy weapon systems.

3-5.6. SEMICONDUCTOR BRIDGE. The semiconductor bridge (SCB) is in production in DOD and DOE programs, as well as in commercial applications for oil well case perforation, seismic charge detonation, and mining applications where precise timing is required. The SCB is essentially a polysilicon resistor on a silicon or sapphire substrate. The SCB method uses a polysilicon bridge that is much smaller than conventional bridgewires and can be manufactured for surface mount technology; allowing other components such as transient suppressors to be integrated on the die with the bridge. The typical bridge is approximately $100 \times 400 \times 2\ \mu m$ and can have resistance values (by design) in any range; however, they are typically 1 to 2 ohms. Passage of a current pulse, with significantly less energy than required for hot bridgewire ignition, generates a plasma discharge in the SCB with a temperature of approximately 5000 degrees C, which ignites the energetic material pressed against the bridge and produces an explosive output in a few microseconds. In contrast to the carbon bridge, the electrical characteristics of the SCB (e.g., bridge resistance, firing energy, etc.) display minimal variation from device to device. SCB's are typically initiated with a short-duration, low-energy pulse (usually from a capacitor) that causes the bridge to heat to the point of plasma formation. The thermal energy of the plasma is transferred to the explosive, thereby initiating the device. Although the no-fire energy is low, the power required to initiate the SCB (in this short-pulse mode) is on the order of several hundred watts. SCB's are thermally fast devices with estimated time constants ranging from 10 of 100 μs. However, HERO testing has shown that the SCB's without additional transient suppression are at least as sensitive to dudding by EM radiation generated arcs as conventional bridge-type devices.

3-5.7. SEMICONDUCTOR IGNITOR. The Semiconductor Ignitor (SCI) is a solid-state initiator developed primarily for application in the M52A3B1 electric primer (conductive mix) used in 20 mm PHALANX ammunition. The development of this technology was prompted by the high sensitivity (and consequent susceptibility) of the M52A3B1 to electrical stimuli, including RF energy. The SCI is comprised of back-to-back diodes (either Schottky-barrier diodes or p-n junctions), one of which is used to convert the electrical energy in the PHALANX firing signal to thermal energy. This thermal stimulus is used to ignite a primary explosive that is in direct contact with the diode surface. The diodes are also designed to *block* low-voltage direct-current (DC) and low-frequency signals, providing a measure of protection from extraneous signals. These devices are also thermally fast.

3-5.8. LASER DIODE INITIATED DEVICES. Laser-ignited explosive/pyrotechnic components can use a rod or semiconductor laser to produce sufficient energy through an optical fiber to ignite energetic materials. Laser ignition of energetic materials is mainly a thermal phenomenon where the material absorbs the incident radiation and ignites when it reaches its autoignition temperature. Therefore, the energy required to ignite energetic materials is a function of many parameters, including thermal conductivity, absorption spectra, composition, packing density, particle size, powder additives or dopants, and blend homogeneity. In order to improve the optical absorption of laser energy, it is necessary to add dopants; usually carbon black, graphite, or infrared absorbing dyes to several of the energetic materials. The physical appearance of the powder cavity in laser-fired devices is similar to their electrical counterparts. The laser energy can be transmitted either through a fiber pin or transparent hermetically sealed window. Laser-ignited fiber pin components are fabricated using short lengths of optical fibers which function in the same general manner as the metal pins of EED's.

Window devices are classified as components that contain a transparent medium or window that are hermetically sealed within the structural member or shell. The window acts as a transparent bulkhead between the optical fiber from the laser and the energetic powder. Since the diameters of window components are typically many times larger than fiber pin components, there are minimal signal losses caused by misalignments between the optical fiber and the window. However, because the window does not act as an optical waveguide, the incident light diverges as it travels through the thickness of the window; requiring high-power sources or more highly doped explosives. Although many claims as to their relative insensitivity to EM radiation have been made, no published data supporting these claims are known at the time of this writing. The properties of the materials used to dope laser-ignited explosives may lend these devices susceptible to coaxial mode current discharges to the components case. In addition, there exists the potential of EM interference in the laser triggering circuits, causing them to function as designed; that is, firing the device.

3-6. SENSITIVITY MEASUREMENTS.

In addition to selecting an EED of a suitable type, the designer must know its firing sensitivity (see table 3-1). The various types of EED's now available are usually classified by their sensitivity to a normally applied current in the differential mode of excitation. Maximum No-Fire Stimulus has been established as "the greatest firing stimulus which does not cause initiation, within five minutes, of more than 0.1 percent of all electric initiators of a given design, at a 95 percent confidence level."

The statistical test commonly used to determine current sensitivity is the Bruceton Test. This test yields an excellent estimate of the mean, but a poor estimate of the standard deviation. When an EED supplier or manufacturer gives no-fire characteristics, the weapon designer should determine what method was used to obtain these characteristics before they are accepted.

In general, the designer is given a requirement for an EED which will perform a certain function within a specified time after the application of the firing stimulus. Also, a certain reliability requirement is attached to the performance of this function. The obvious approach to fulfilling these requirements is to use the largest power source allowed by the system in conjunction with a sensitive EED. However, in designing with HERO in mind, the least sensitive EED should be used.

Table 3-1. Types of EED's and Typical Characteristics

ELEMENT TYPE	DC RESIST. (OHMS)	SENSITIVITY		FUNCTION TIME
		NO-FIRE	ALL-FIRE	
HOT BW (standard)	0.1 – 10	$0.1A$	$1.0A$	$\mu s - ms$
BW 1 Amp/1 Watt	1.0	$1.0A$	$5.0A$	ms
SEMICONDUCTOR	0.05 – 1.0	$1.0A$	$10A$	ms
EBW (gaped)	∞	$0.1\mu F/300V$	$0.1\mu F/1000V$	$<6\mu s$
EBW (ungaped)	0.01 – 0.2	$0/1\mu F/300V$	$0.1\mu F/2000V$	$<6\mu s$
EFI	0.01 – 0.1	$0.25\mu F/1000V$	$0.25\mu F/1500V$	$<2\mu s$
CARBON BRIDGE	800 – 12000	$10V$	$1000V$	$<1\mu s$
CONDUCTIVE MIX	$0.01 - 10^6$	$0.1A$	$10A$	μs
SPARK GAP	∞	$150V$	$300V$	μs

CHAPTER 4

FIRING SYSTEM DESIGN

4-1. GENERAL.

Many of the weapon systems that are used by the Navy have subsystems and firing systems that are exterior to the ordnance itself. These exterior systems, together with their connecting circuitry, can exacerbate the HERO problem. The design of ordnance that meet the HERO requirements requires that the effect of the electromagnetic environment (EME) on the system as a whole be considered for all situations that the system is expected to encounter in its stockpile-to-safe separation sequence. There are many situations during this sequence when electromagnetic (EM) energy can enter the ordnance. This energy must be excluded or its effects mitigated at all times if the weapon system is to be considered adequately designed. In addition, the weapon system must be designed so that the handling, loading, and testing techniques that must be used do not create additional HERO problems.

The design of the firing system is of particular importance in reducing the susceptibility of the ordnance to the EME. Because the firing system provides the path for transferring the firing energy to the electrically initiated device (EID), it can also provide the path for transferring EM energy to the EID. Only in a few types of ordnance will the firing circuit be completely contained within the structure so that the required level of shielding effectiveness is provided by the metallic skin. When the level of shielding effectiveness provided by the system is not sufficient to preclude HERO, the designer will need to use the firing system design practices discussed in this chapter.

4-2. FIRING SYSTEMS.

A firing system, for the purpose of this discussion, consists of a power source, transmission lines, and all control and switching circuits required to control and transfer power to an EID. Figure 4-1 illustrates the basic elements of a typical firing system. All firing systems can be divided into two basic categories: low voltage systems used to initiate low energy devices such as EED's, and high voltage systems used to initiate exploding bridgewire (EBW) devices and exploding foil initiators (EFI's). There are many variations of these two types. Firing techniques can vary from a simple switch closure to sophisticated coded-pulse systems. The exact nature of the mechanism used to initiate the EID in any type of ordnance is usually dictated by the mission or specific ordnance application.

Figure 4-1. Basic Elements of Firing Systems

EM sources with frequencies above 10 KHz should not be used to provide the initiating energy for EID's. If a frequency-coded firing system is used, the receiving equipment as well as the firing system must be protected from the EME. The receiving equipment must not permit false indications during exposure to the environment since this might result in premature EID initiation and possible ordnance actuation.

4-3. FIRING SYSTEM DESIGN PRACTICE.

Poor wiring practices are prime factors contributing to the coupling of EM energy into a firing system. Among the areas in which this commonly occurs are circuit configuration and cable routing. Figure 4-2 illustrates poor wiring techniques from the HERO standpoint. The rocket launch tube is insulated from the rocket launcher pod and serves as one of the firing contacts. One lead of the EID is connected to the weapon skin and hence to the launch tube by a contact spring when the weapon is loaded. The other side of the EID is brought out of the weapon to a firing button, which is electrically connected to the launcher pod.

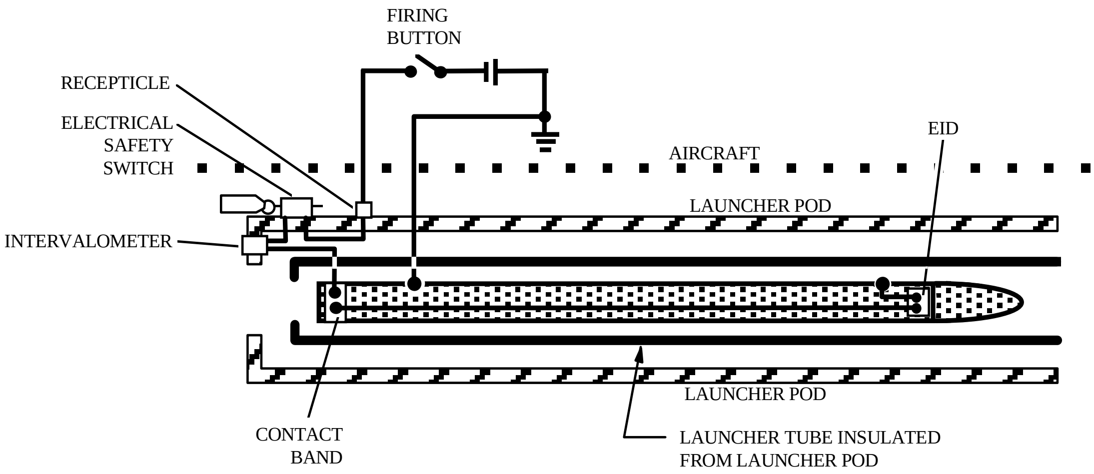
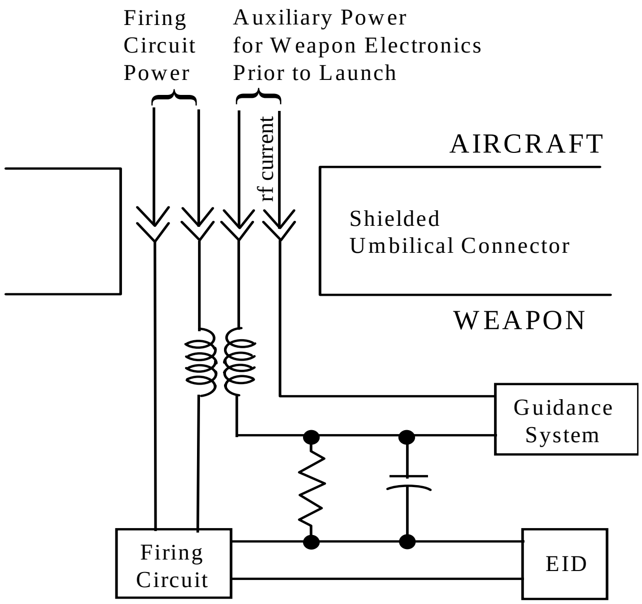


Figure 4-2. Improper Firing System Wiring

This configuration is particularly susceptible during any handling and loading operation. If personnel touch the weapon skin after the firing leads are connected, EM energy can be coupled from the aircraft through the EID to the deck. This firing circuit design is basically hazardous. If the weapon were to be made immune to RF energy, the firing circuit would have to be redesigned. Firing circuits should always be a two-wire balanced system isolated from ground so that no direct path for EM energy to the EID exists during handling or loading of the weapon.

Improper routing of firing circuit wiring or cables can cause the weapon to become susceptible to RF energy. All firing circuit wiring should be electrically isolated from other wiring and cables in the system to prevent coupling energy from one circuit to the other.



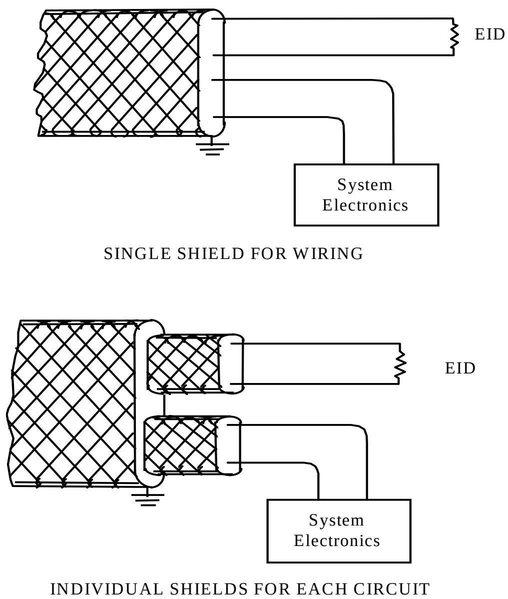


Figure 4-3. Mutual Coupling Between Cables

Figure 4-4. Single Common Shield and Individual Shields

Coupling between circuits exists when the current flowing in one circuit produces a current in the other. The mutual elements that can couple energy are resistance, inductance, capacitance, or any series or parallel combination of these elements. An example of coupling possibilities is suggested in figure 4-3. Coupling can be reduced by shielding each circuit, or by the physical separation of the wiring (see figure 4-4).

To prevent energy from the EME from coupling into the wiring within a shielded enclosure, circuit conductors shall not penetrate holes in the shield unless shielded and terminated as described in chapter 5. Also, conductors shall not pass within 1-inch of holes in the shield, and these holes shall be no greater than ¼-inch in diameter unless screwed or protected by waveguide attenuator. This is illustrated in figure 4-5.

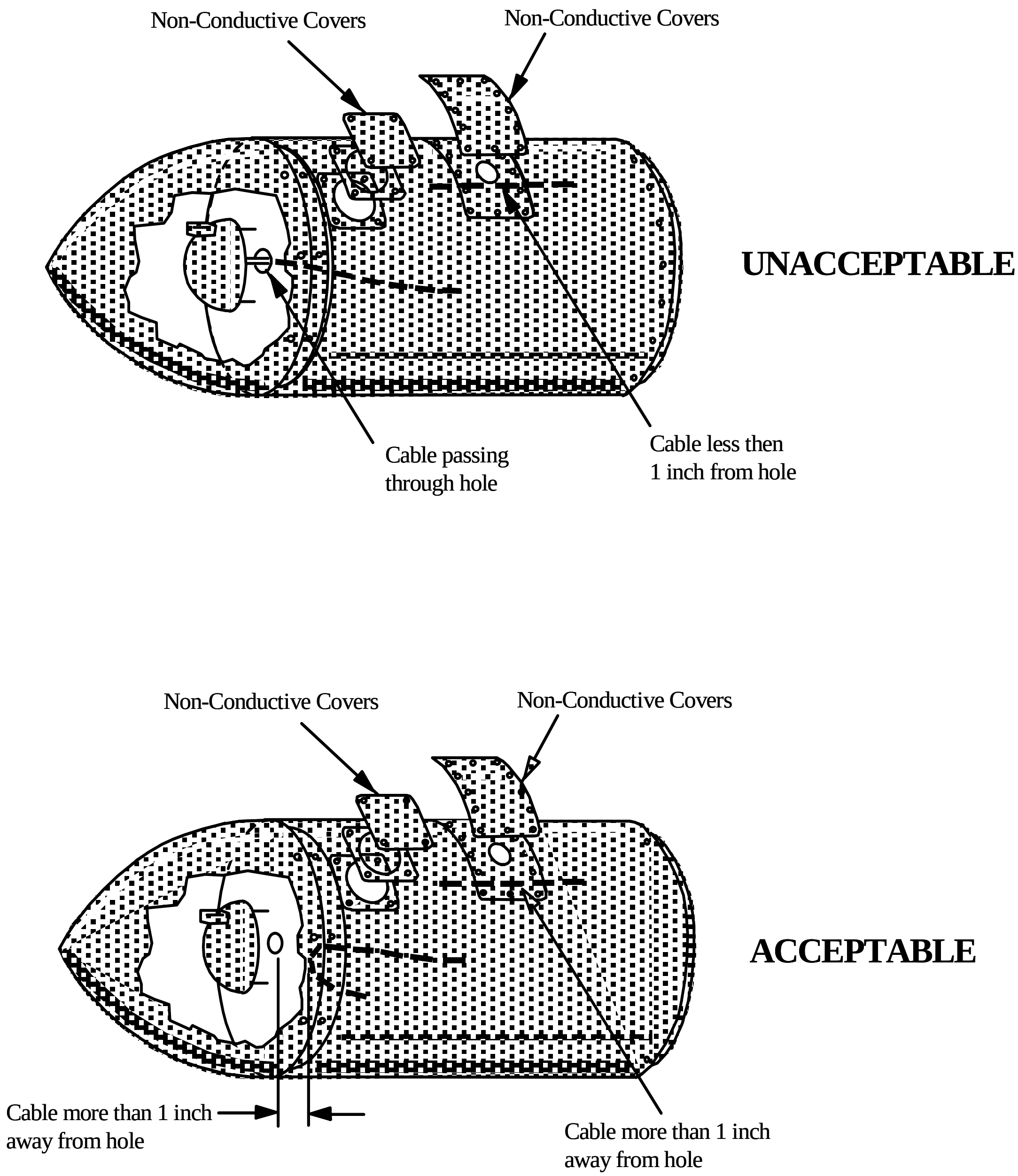


Figure 4-5. Holes in Partially Shielded Weapon Sections

EID firing circuit wiring should be as short as possible and the leads equal in length to minimize induced voltages, as shown in figure 4-6. The firing circuit leads should be twisted uniformly to reduce the effective area of the pickup loop created by them and to cancel the voltage that may be induced.

Caps or shorting plugs are required during storage on many weapons for protection from static charges. From the HERO standpoint, caps are preferred to shorting plugs because a cap has no actual connection to the firing circuit. The cap should be conductive so that it completes the shield when it is installed. In some cases, a shorting plug can actually increase the susceptibility of the weapon to EM energy by creating a loop antenna with the EID circuit. Also, during its removal and replacement, it can provide a path for radio frequency (RF) currents to flow to the EID circuit. Shorting plugs can be designed so that they reduce the HERO problem (figure 4-7). They must be constructed of conductive material and designed so that during installation, the shield makes and maintains peripheral shielding contact prior to the shorting of the firing circuit. Also, a good insulating coating on all exposed surfaces of the plug will add additional protection during installation and removal. The thicker the insulation (up to approximately ¼-inch), the more protection it will provide, particularly in the 2 to 32 MHz frequency range.

Some multistage weapons require exhaust ports for venting the engine exhaust generated during stage separation. These exhaust ports are permanent apertures in the weapon shield; therefore, all cabling and components of the firing system in this system of the weapon must be carefully shielded and filtered to preclude HERO.

If sections of the weapon are non-metallic, all cables and wiring in the firing system that pass through this section must be properly shielded and filtered to preclude the HERO problem. Non-metallic housings, such as fiberglass, plastics and composite materials, do not afford any HERO protection to the firing system. Metallic coatings can be applied to non-metallic structures to help maintain shielding integrity.

4-4. SAFE AND ARM DEVICES.

Figure 4-8 is an example of a typical hot bridgewire (HBW) firing system that includes an electrical safe and arm device. In this example, the firing leads between the power source and the EID are opened, and the EID leads are shorted to ground by the safe and arm switch. The open contacts in a properly filtered firing system will provide protection from arcs that might cause initiation during the loading and handling operations. The arming process causes the switch sections to move, removing the short to ground from the EID leads and connecting the EID leads to the firing circuit. The ordnance is thereby armed and ready for firing.

A mechanical safe and arm device such as shown in figure 4-9 is often used to misalign the explosive train when in the safe condition. It does not, however, solve the HERO problem because the EID is not affected either mechanically or electrically by the functioning of the device. Thus, the EID can still be inadvertently initiated or dudded by the EM energy. This type of device is used primarily for safety reasons and is often combined with the electrical safe and arm device.

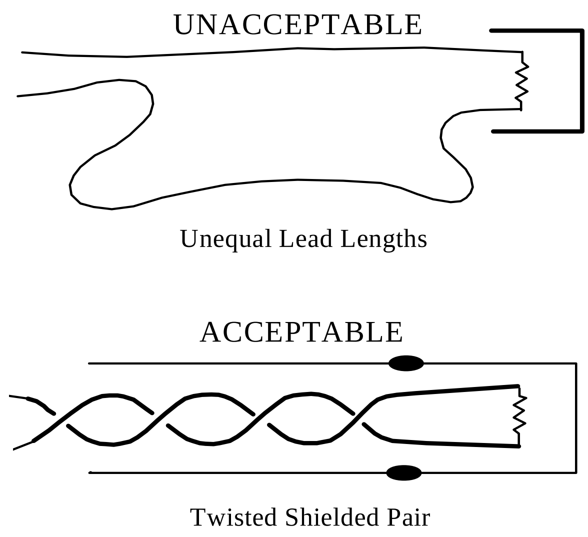


Figure 4-6. Unequal Lead Lengths and Twisted Shielded Pair

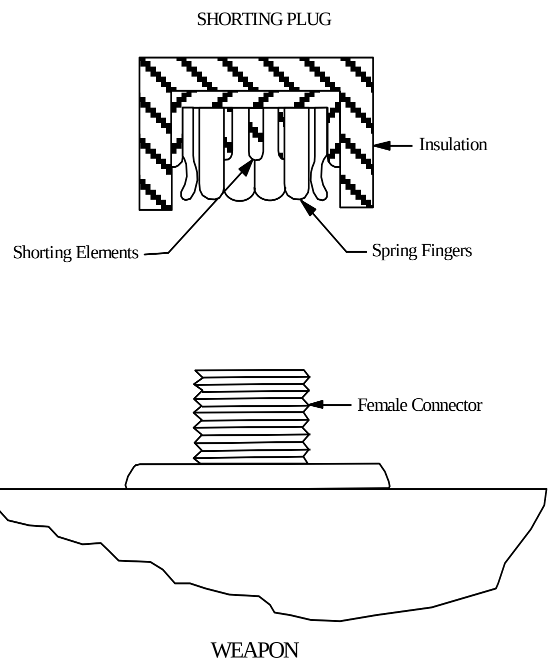


Figure 4-7. Shorting Plug for Weapon

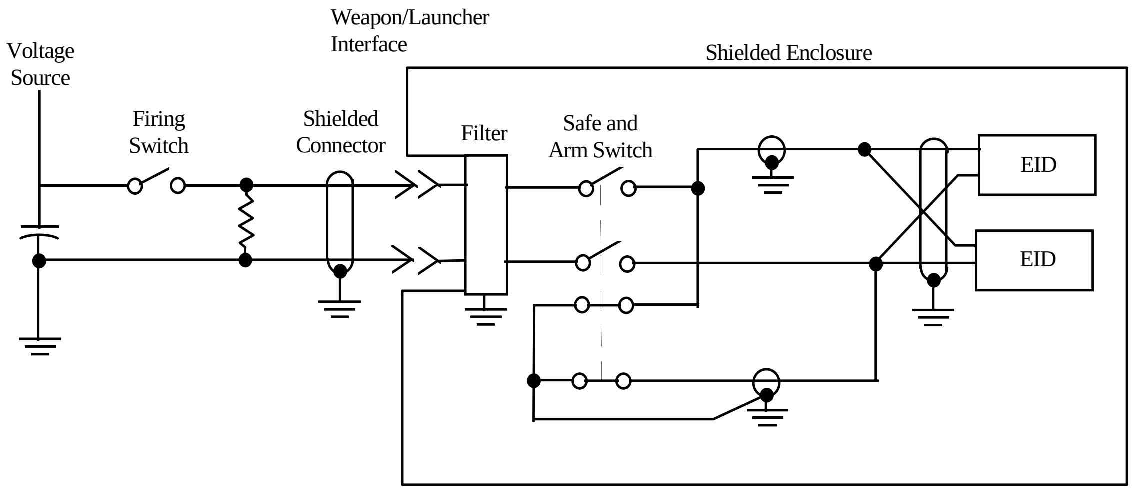


Figure 4-8. Typical Hot Bridgewire Firing Circuit and Safe and Arm Device

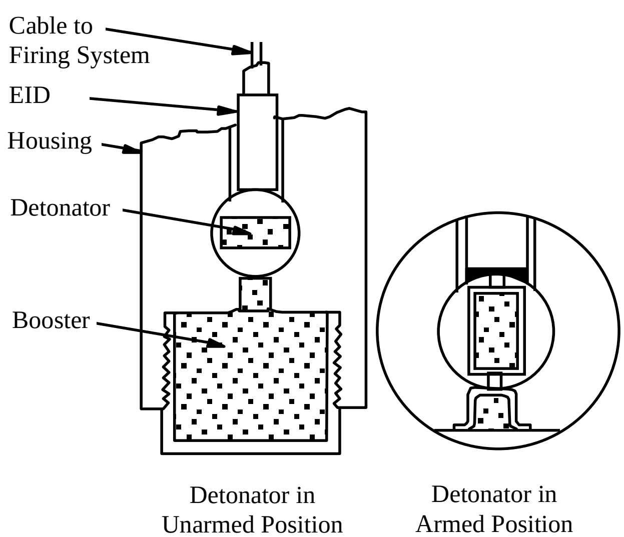


Figure 4-9. Mechanical Safe and Arm Device

4-5. HERO PROBLEMS OF FIRING SYSTEMS.

Examples of firing system designs that cause ordnance to be HERO SUSCEPTIBLE are given in this section. These examples are based on actual weapon design, and the expedients discussed are considered interim measures (retrofits) allowing ordnance to remain operational in present EME's. They are not considered as having completely solved the HERO problem or as having rendered unsatisfactory designs completely satisfactory.

Aircraft and surface-launched weapons pose the greatest hazard because they must be handled and loaded in high-level EME’s, and they generally have subsystems or firing systems that are exterior to the weapon. Underwater-launched weapons are not usually exposed to the high-level EME’s, and the nature of their design is such as to provide more protection from the EME than is provided by either air or surface-launched weapons. They can be exposed, however, to high-level EME’s, particularly when they are being transferred to a ship or submarine.

Figure 4-10 shows a typical aircraft weapon firing system. As can be seen, the cables attached to the weapon can be quite extensive as they thread through the aircraft. The cables run from the pilot's control console (1) through the fuselage, adjacent to radio and radar equipment (2), into a multi-conductor cable bundle (3), through the wing panel in a cable bundle (4), through the pylon and launcher (5), then to the weapon igniter (6). The cable can have a length of about 25 feet, and can be a very effective antenna in an EM field.

Figure 4-11 shows an air-launched weapon in which the connection from the aircraft firing system to the weapon is made through button contacts. These button contacts make the weapon particularly susceptible to HERO during the handling and loading procedures because personnel can touch the contacts and conduct EM energy directly into the EID. This method of connecting the aircraft firing system should be designed in such a way as to prevent personnel or tools from touching the conductors that lead into the weapon.

Figure 4-12 shows an air-launched weapon partially loaded into its launcher. This weapon is connected electrically to the aircraft firing system by contact rings. These rings are exposed during handling and loading, thus EM energy can be coupled into the weapon and make it susceptible. To help reduce the hazard, a removable shielding band, as shown in the figure, was designed to cover the exposed contacts. This is not a completely satisfactory solution since the bands are not an integral part of the weapon. Also, the weapon requires elaborate handling and loading procedures since the bands are removed during loading and the weapon can be unloaded without them, leaving it susceptible. Elaborate handling and loading procedures should not be relied upon to solve the HERO problem, because failure to implement them will create a HERO problem.

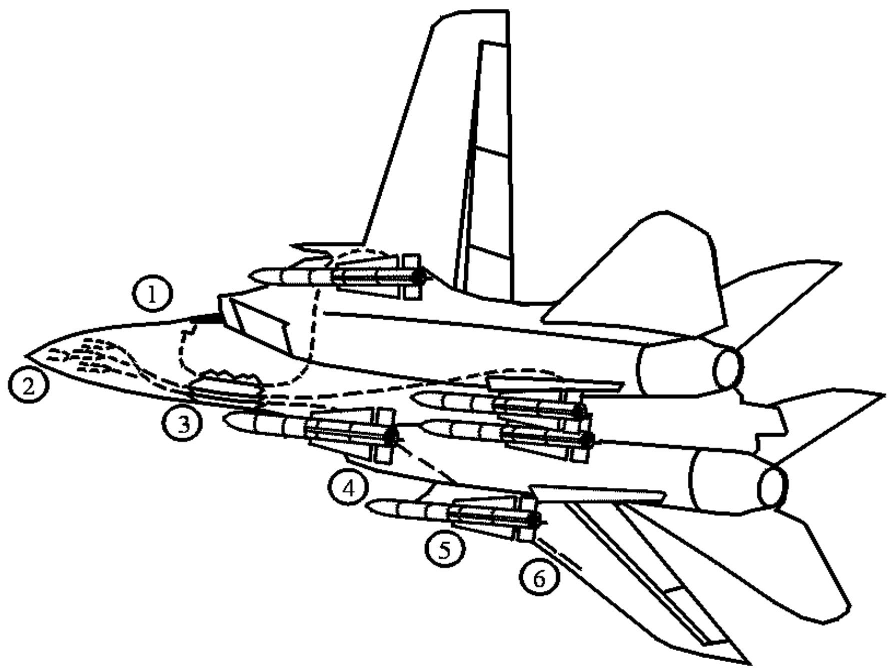


Figure 4-10. Typical Aircraft Weapon Firing System

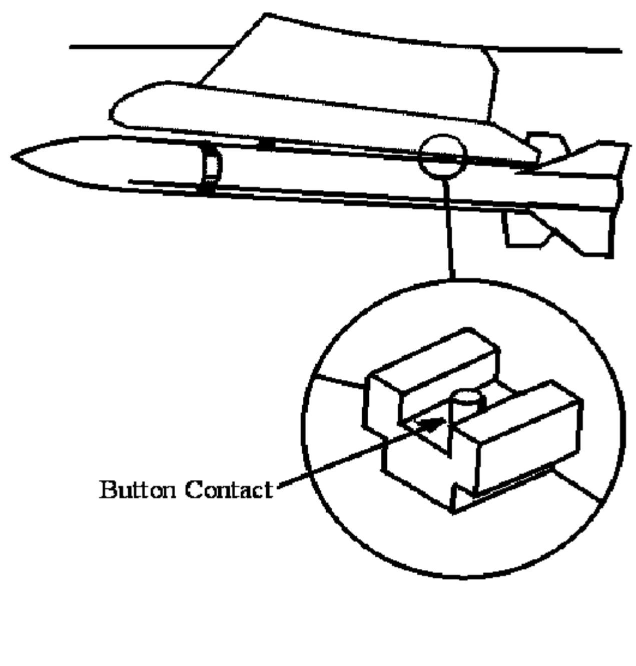


Figure 4-11. Air-Launched Weapon

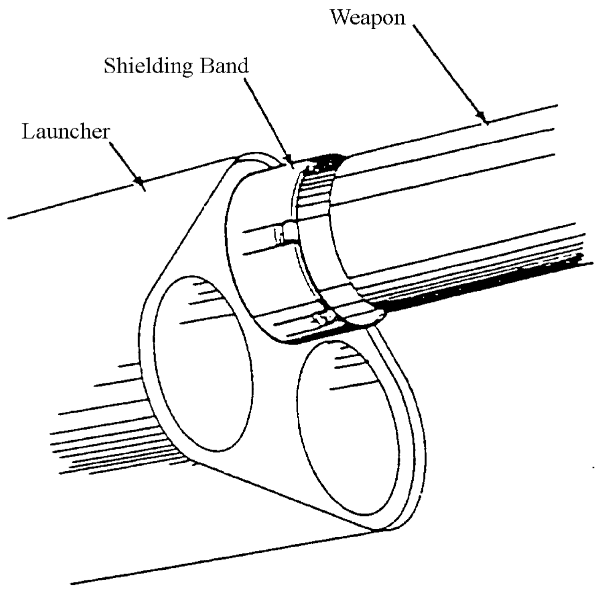


Figure 4-12. Air-Launched Weapon Partially Loaded into Its Launcher

Figure 4-13 shows a weapon/launcher interface. In the illustration, the umbilical cable is being connected before the weapon is racked to the launcher. This can be a hazardous situation because in an EME the launcher and the weapon can be at a different RF potential. This difference in potential can cause a flow of RF current in the weapon and greatly increases the possibility of generating an arc as the umbilical is being connected. It may not always be obvious that a high-potential difference can exist between aircraft and deck. However, near a vertical whip antenna radiating in the 2 to 32 MHz frequency range, a potential difference of greater than 500 volts can exist between aircraft and deck, even if conductive tie-downs are used.

After the weapon has been secured to the launcher, the RF potentials on the launcher and the weapon are the same or nearly the same, and the possibility of large RF currents and arcs is greatly reduced. Therefore, the weapon should be designed so that it must be racked before the umbilical cables can be connected. This is illustrated in figure 4-14. The umbilical cable should be as short as practical, and the cable and its connectors should be completely shielded with the shields properly bonded around the periphery of the connector. The connector should be designed and installed with the male portion on the umbilical cable and the female portion on the weapon. The contacts should be recessed and designed in such a way that the connector shield mates before the connector pins mate.

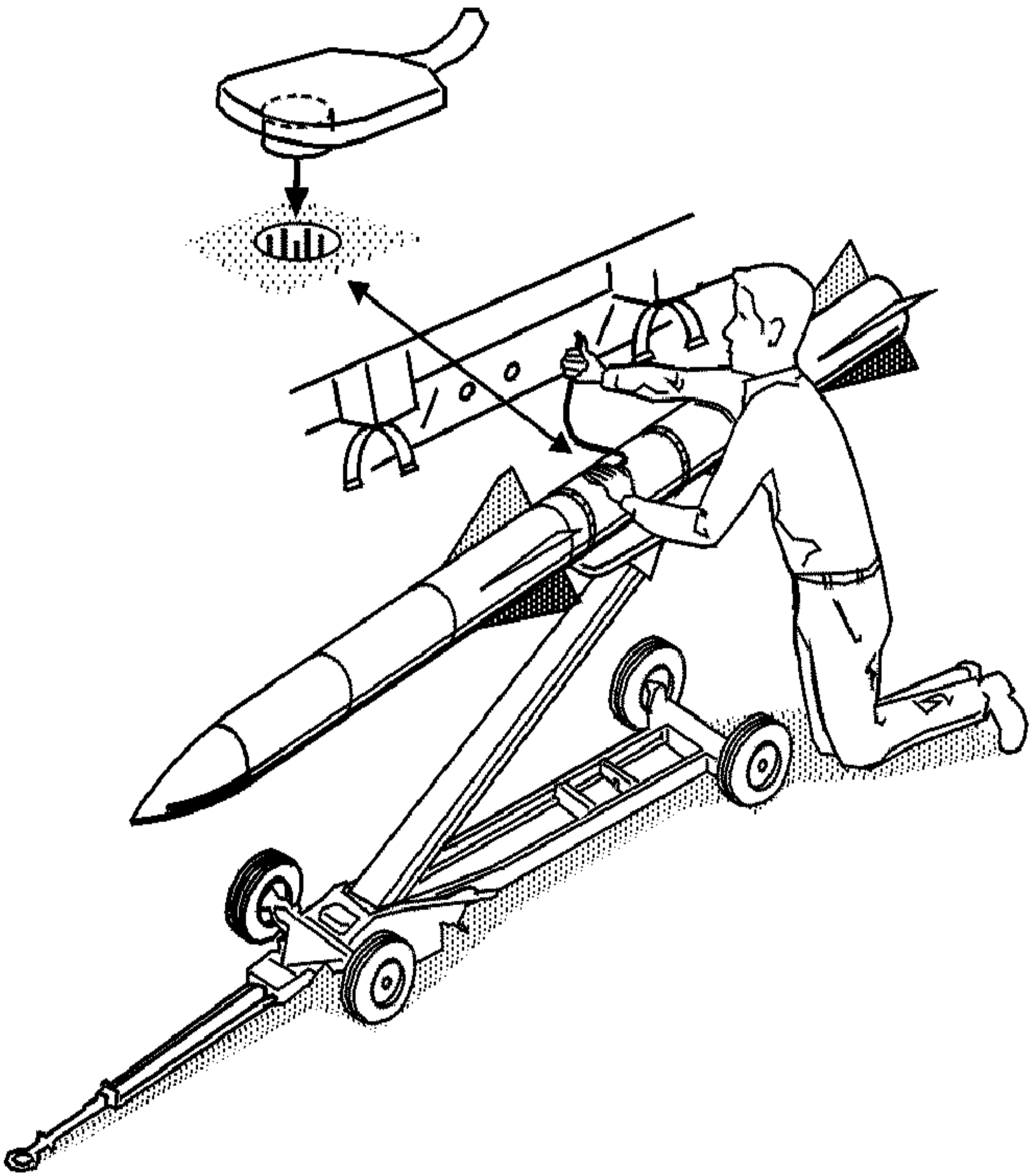


Figure 4-13. Weapon/Launcher Interface/and Umbilical Mating

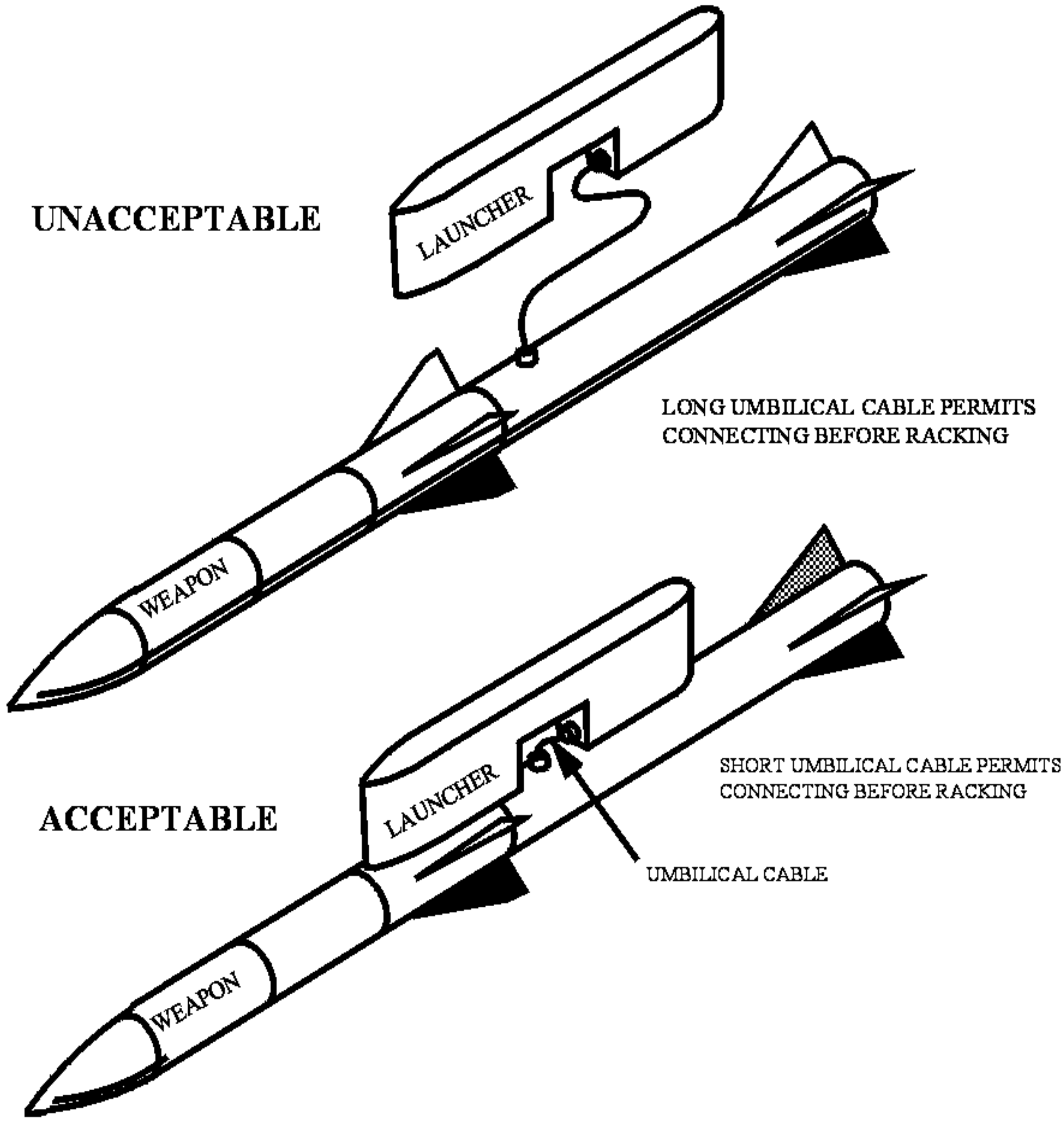


Figure 4-14. Unacceptable Method of Umbilical Mating

Figure 4-15 shows a test set being used to test a weapon station on an aircraft. This can be a hazardous operation because the test set and external cables can couple energy into the firing system of the aircraft. This energy can be conducted to weapons already loaded. The design of the test equipment and its cabling must be given the same consideration as the design of the weapon itself, if a HERO problem is to be prevented. Also, should EM energy enter the test equipment, it may give false indications. Therefore, the cable and connectors must be properly shielded with the shield bonded to the weapon and to the test equipment to preserve the integrity of the weapon enclosure.

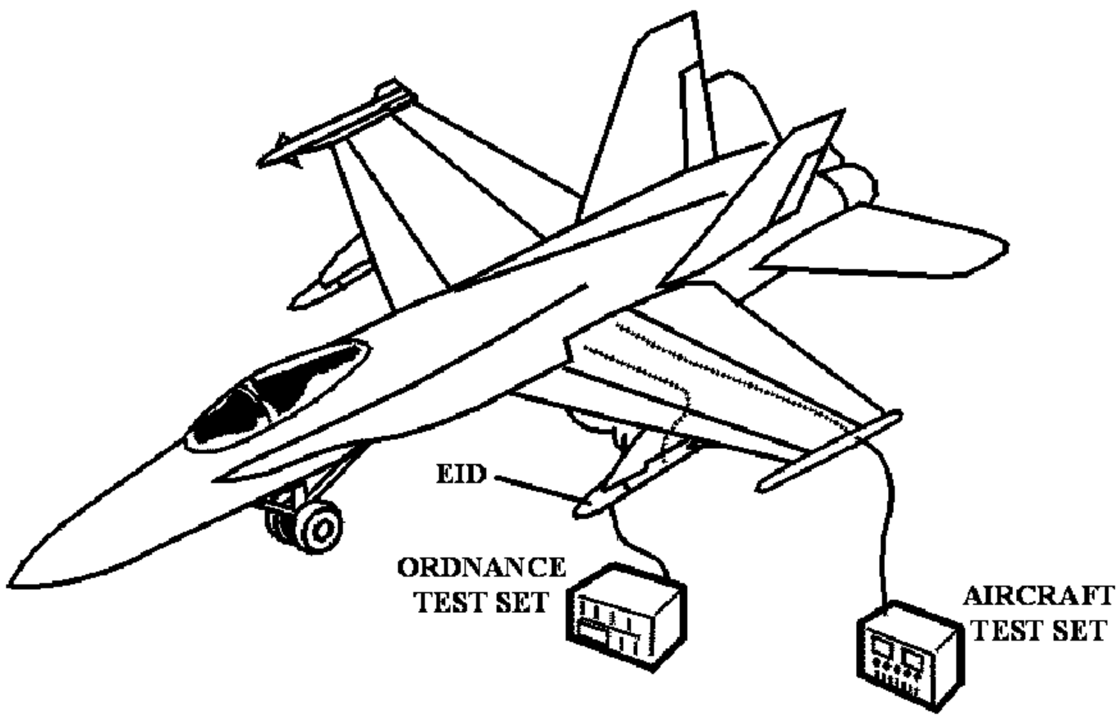


Figure 4-15. Launcher/Test Set Interface

Figure 4-16 shows a typical surface-launched missile with its launcher and control cables. The cable runs from the fire control panel (1) in an armored multi-conductor missile control and monitor cable bundle (2), through transfer panels (3) to a slip ring assembly in the launcher pedestal (4), and emerges from the launcher to contact firing shoes of the missile (5). The total length is approximately 90 feet. Because the fire control, monitoring equipment, and cabling are almost entirely enclosed within the ship's structure, they are protected to some extent. The missile launcher and the umbilical cable, on the other hand, are exposed to the EME. Here again, the missile and the umbilical cable should be properly shielded.

Figure 4-17 shows a surface-launched target with umbilical cables connected to the target and access doors open. The two long umbilical cables create a potential hazard because the two cables can form a loop antenna or the long cables can act as very effective antennas. The number of umbilical cables should be kept to a minimum and the cables should always be as short as possible. The use of access doors or ports may create a hazardous condition because energy can be coupled through them to the firing circuit. When it is necessary to have an access door, all of the cables of the firing system that are exposed when the door is open must be shielded. Access doors should be kept to a minimum.

The ability to shield effectively can be greatly impaired while the target is being prepared for launch. Cables are being handled, connectors are being mated, and access ports on the target may be open. Personnel operating, handling, and loading equipment may, through touch, contribute to the coupling of RF energy into the target. When personnel or equipment make contact with any part of the target, a situation of RF energy transfer could develop that may not have been considered in the design of the target.

Figure 4-18 shows a weapon being lowered through the hatch of a ship by a crane. The handling crane, acting as a receiving antenna, conducts EM energy to the weapon and its shipping container. If the weapon is transferred to the ship in a partially assembled or susceptible condition, or if there is exposed wiring, the shipping crate should be of all-metal construction and should completely enclose the weapon in such a way as to provide a shield during storage and handling. In some cases, such as in underwater-launched weapons, the weapon is transferred to the ship or submarine in an all-up condition without a container. Care must be taken by the designer to assure that the weapon is HERO SAFE during this operation.

Figure 4-19 shows the loading of a fuze in a weapon. There is a specific hazard to personnel during this operation. This operation must be performed in an area free of EM energy. If loading a fuze or performing maintenance operations in the EME is required, the exploder should be completely shielded. The cables and connectors should be designed in such a way as to preclude arcing and the entry of EM energy. In addition, the operations must undergo an evaluation test to verify that no potential HERO problems exist.

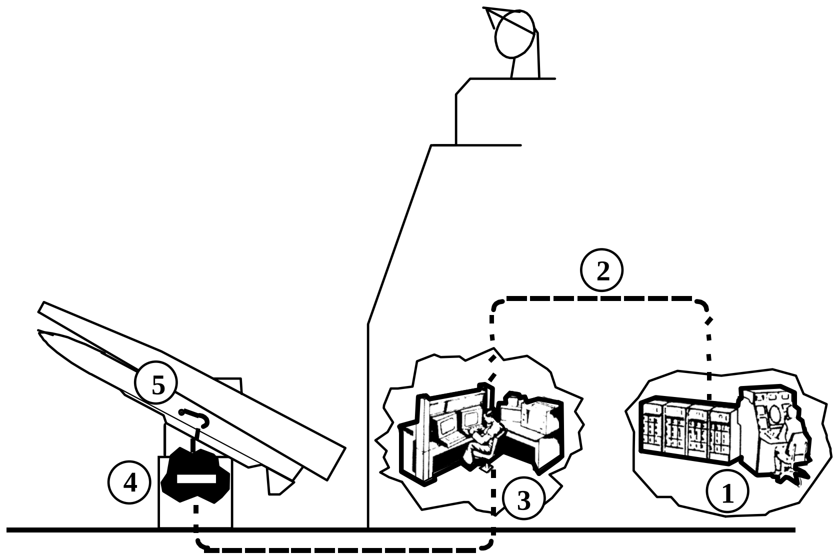


Figure 4-16. Typical Surface-Launched Missile System

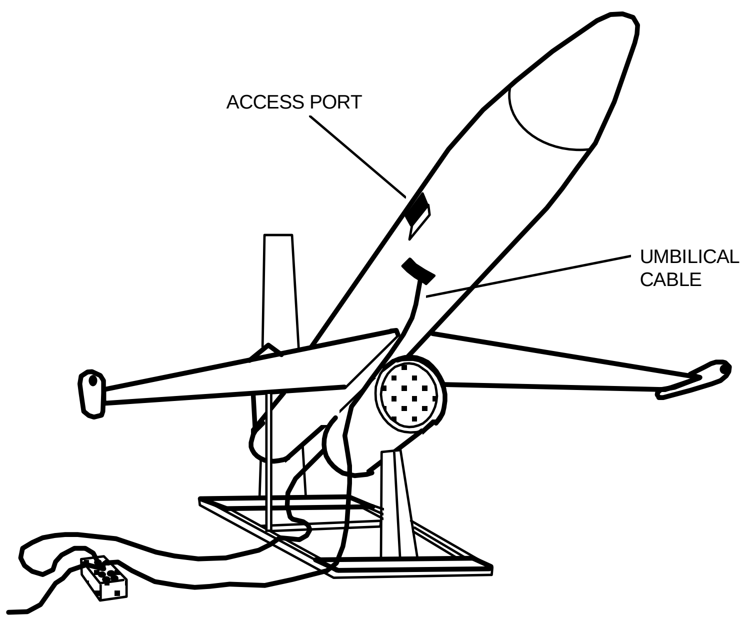


Figure 4-17. Surface-Launched Target

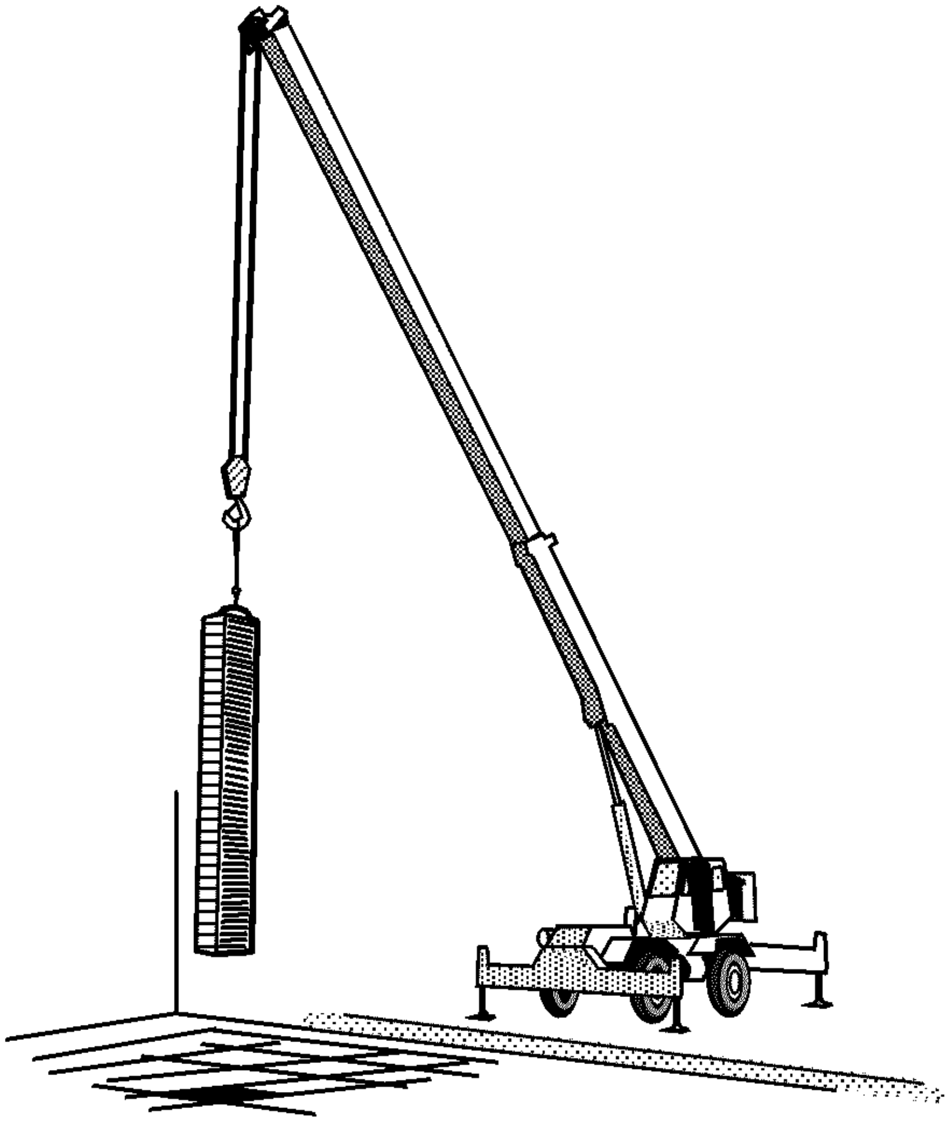


Figure 4-18. Weapon Being Lowered through Ship Hatch

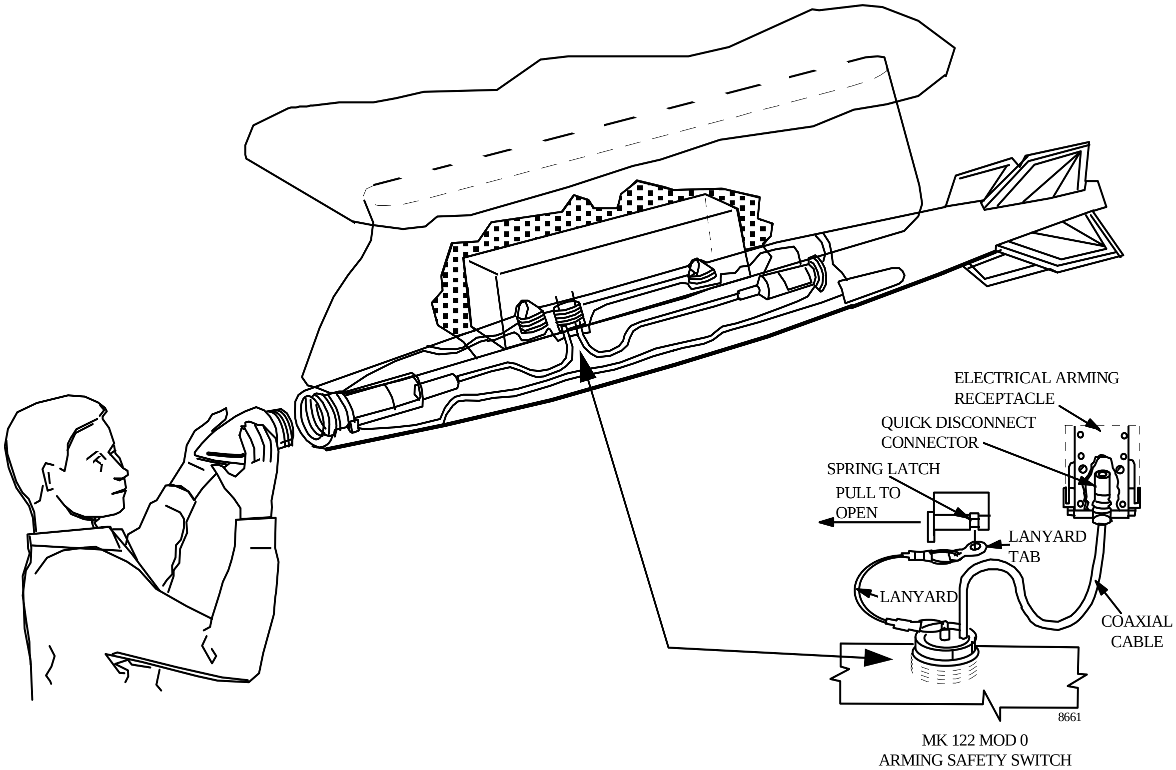


Figure 4-19. Loading of a Fuze in a Weapon

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CHAPTER 5

SHIELDING

5-1. GENERAL.

The practical approach for solving the HERO problem is to provide a complete radio-frequency (RF) shield for the electrically initiated devices (EID's). If it were not for the many mechanical and electrical interfaces required in weapon systems, the shielding problem would be reduced to choosing a proper shield material and applying simple shielded box concepts. However, since interfaces do exist and must be accommodated, the selection and implementation of techniques to provide continuity at these interfaces become an integral part of the weapon system design to mitigate the HERO problem. The selection of materials to provide shielding and the techniques to ensure shielding integrity are described in this chapter. Figures 5-1 and 5-2 illustrate some of these interfaces and indicate the range of proper and improper shielding design techniques.

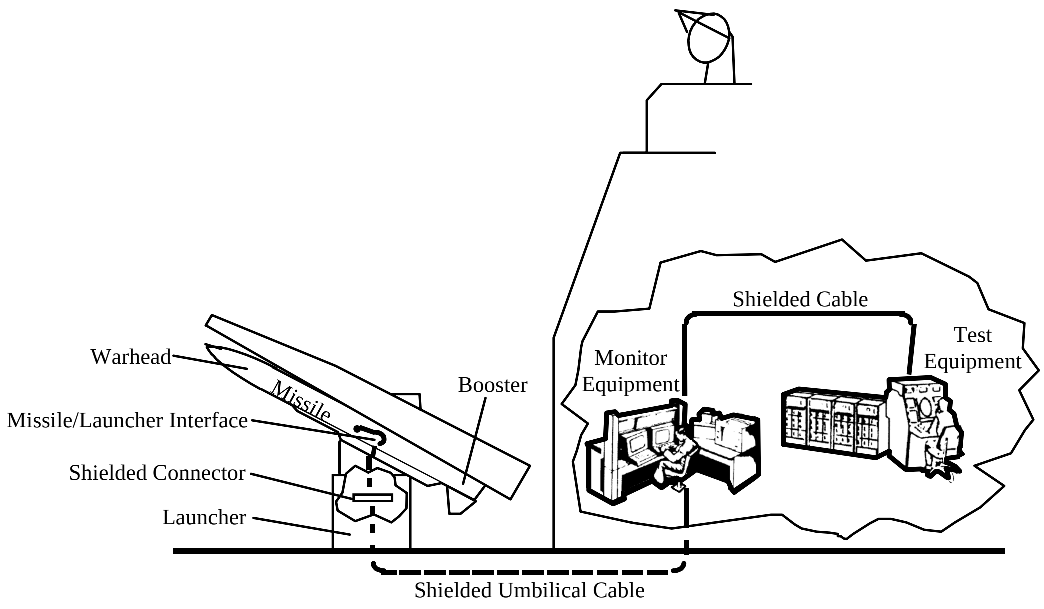


Figure 5-1. Typical Missile System Shielding Interfaces

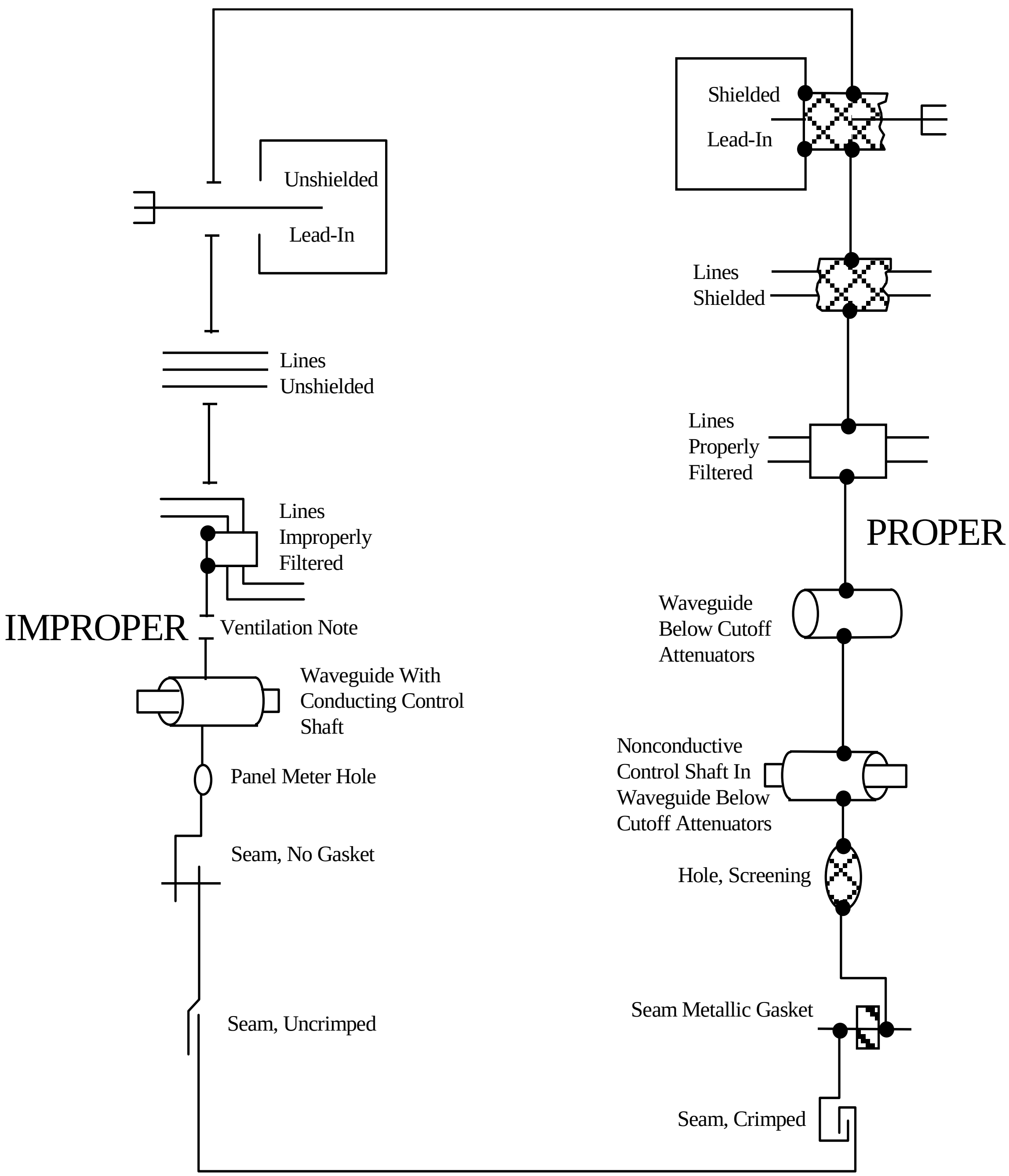


Figure 5-2. Typical Shielded Compartment Discontinuities-Proper and Improper

5-2. GENERAL SHIELD DESIGN CONSIDERATIONS.

Shielding has two main purposes: to keep radiated electromagnetic (EM) energy confined within a specific region, and to prevent radiated EM energy from entering a specific region. Thus, shielding is essentially a decoupling mechanism used to reduce radiated interactions between equipments, or between portions of a given equipment.

Shielding requirements, as far as representative weapon system design is concerned, are of greatest importance with regard to shipboard operations. A major consideration in the development of weapon systems is the design of shielding to protect the ordnance from the expected high-level electromagnetic environment (EME).

The shielding effectiveness of a weapon system is a complex function of a number of parameters, the most notable of these being the frequency and impedance of the impinging wave, intrinsic characteristics of the shield materials, and the numbers and shapes of shield discontinuities. The design process consists of establishing undesired EME levels on one side of a proposed shielding barrier, estimating tolerable EME levels on the other side, and trading off shield design options to achieve the necessary shielding effectiveness level.

This chapter identifies many of these shield design options. It is subdivided into discussions of solid shields (including multiple wall and thin-film shields), non-solid shields, cable shielding, and the maintenance of shielding integrity through a connector, and shield discontinuities (including seams, gasket requirements, the use of waveguide below cutoff attenuators, conductive glass, and other aspects).

Throughout the discussion, it should be recognized that shielding, although an important technique for reducing HERO effects, is not the only technique available for this purpose. Application of shielding techniques should not be made without due regard to the roles that filtering, grounding, and bonding will play in the suppression program.

5-3. SOLID SHIELDING MATERIALS.

5-3.1. SHIELDING ANALYSIS. EME attenuation by a solid shield is due to two distinct effects: (1) reflection (due to impedance mismatches) of the interference wave at the air-metal boundary as the wave strikes the metal surface, and reflection at the metal-air boundary as the interference wave emerges from the metal shield and (2) absorption of the interference wave in passing through the metal shield between the two boundaries. The first loss is generally called Reflection Loss and the second is called Absorption or Penetration Loss. The combined loss due to these two effects is considered the shielding effectiveness of the shield.

It is convenient to separate the initial reflections from both surfaces of the shield from subsequent reflections that may take place at these surfaces. These effects are called the Single Reflection Loss and Multiple Reflection Correction Term, respectively. Under circumstances when the absorption loss is greater than about 15 dB, the multiple reflection correction term can be ignored.

Using transmission line theory, the <u>Shielding Effectiveness</u>, S, of solid shielding materials is defined as:

$$S \ (in \ \ dB) = A + R + B \qquad (5\text{-}1)$$

where

A = Absorption (Penetration) Loss by the material, in dB

R = Single Reflection Loss from both surfaces of the sheet, in dB

B = Multiple Reflection correction term, in dB

Magnetic shielding depends primarily on absorption losses, since reflection losses for magnetic fields are small for most materials. Electric fields are readily attenuated by metal shields because large reflection losses are easily obtained. Absorption loss is the same for electric and magnetic fields.

The absorption loss term, A, is independent of wave impedance, and can be described by the relationship:

$$A \;(in \;\; dB) = 3.34d\sqrt{\mu f g} \qquad (5\text{-}2)$$

where

d = shield thickness, in inches

g = material conductivity, relative to copper

μ = material permeability, relative to copper

f = frequency, in Hertz

For a given material, absorption loss, in dB, at a specific frequency is a linear function of material thickness. The characteristics of the material that influence this loss are its conductivity and permeability. Table 5-1 lists values of conductivity and permeability relative to copper for a variety of useful shielding materials, and indicates calculated values of absorption loss at 150 kHz.

Table 5-1. Characteristics of Various Metals Used for Shields

Metal	Relative Conductivity g	Relative Permeability at 150 kHz u	Absorption Loss db/mil at 150 kHz
Silver	1.05	1	1.32
Copper-Annealed	1.00	1	1.29
Copper-Hard Drawn	.97	1	1.26
Gold	.70	1	1.08
Aluminum	.61	1	1.01
Magnesium	.38	1	.79
Zinc	.29	1	.70
Brass	.26	1	.66
Cadmium	.23	1	.62
Nickel	.20	1	.58
Phosphor-Bronze	.18	1	.55
Iron	.17	1,000	16.90
Tin	.15	1	.50
Steel, SAE 1045	.10	1,000	12.90
Bryllium	.10	1	.41
Lead	.08	1	.36
Hypernick	.06	80,000	88.5*
Monel	.04	1	.26
Mu-Metal	.03	80,000	63.2*
Permalloy	.03	80,000	63.2*
Steel, Stainless	.02	1,000	5.7*

***** Assuming material not saturated.

The single reflection loss term, R, is dependent upon wave-impedance, and can be described by the relationships:

$$R \text{ (low impedance source)} = 20\log\left[\frac{0.462}{r\sqrt{fg/\mu}} + 0.136\sqrt{fg/\mu} + 0.354\right] \quad (5\text{-}3)$$

$$R \text{ (high impedance source)} = 354 - 20\log\left[r\sqrt{\mu^3/g}\right] \quad (5\text{-}4)$$

$$R \text{ (plane wave source)} = 168 - 20\log\sqrt{\mu f/g} \quad (5\text{-}5)$$

where r = source to shield distance, in inches.

The corresponding Nomographs for equations (5-3), (5-4), and (5-5) may be found in figures 5-6/5-7, 5-5, and 5-3/5-4, respectively. These equations are plotted in Figures 5-8 through 5-10. The amplitude of a wave reflected from solid shielding material depends upon the degree of mismatch between the impedance of the impinging wave, the air medium, and the impedance of the shield. It should be noted that the impedance of the impinging wave in the vicinity of the shield will be a function of the distance of the shield from the source and whether or not it is in the near field or far field. This is discussed later in this section.

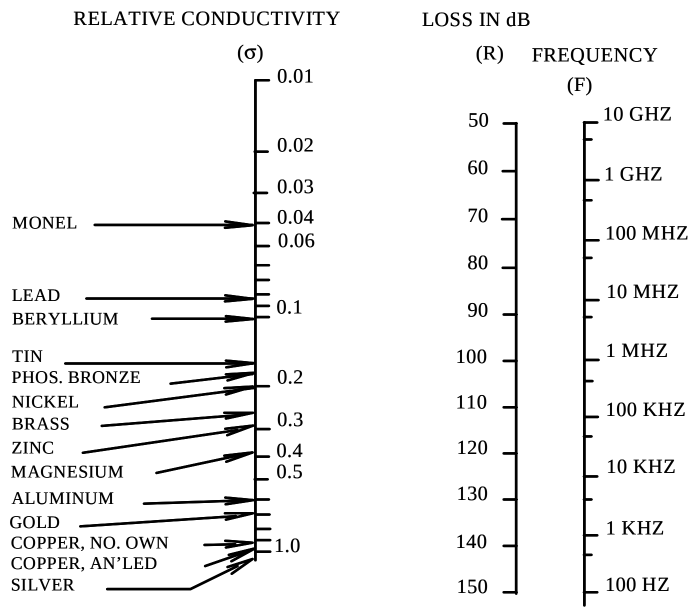


Figure 5-3. Plane Wave Reflection Loss for Nonmagnetic Materials

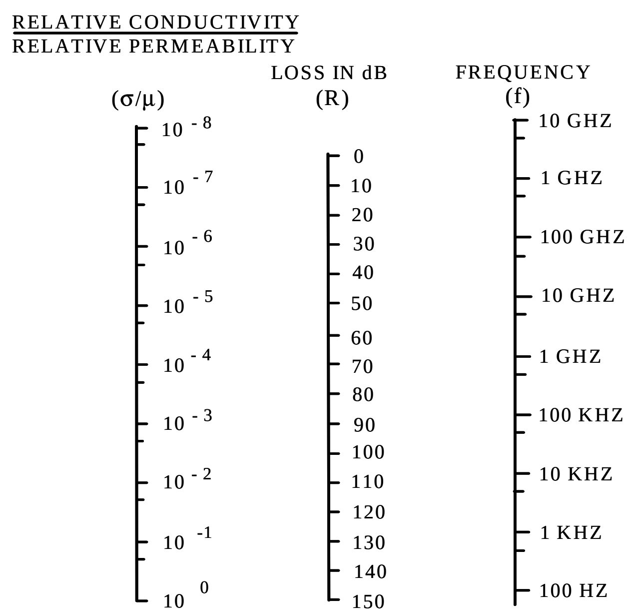


Figure 5-4. Plane Wave Reflection Loss for Magnetic Materials

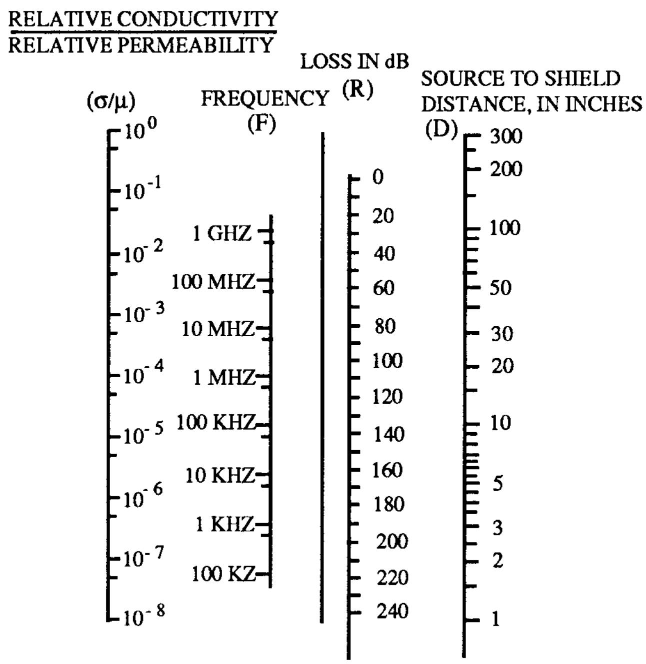


Figure 5-5. Electric Field Reflection Loss for Magnetic Materials

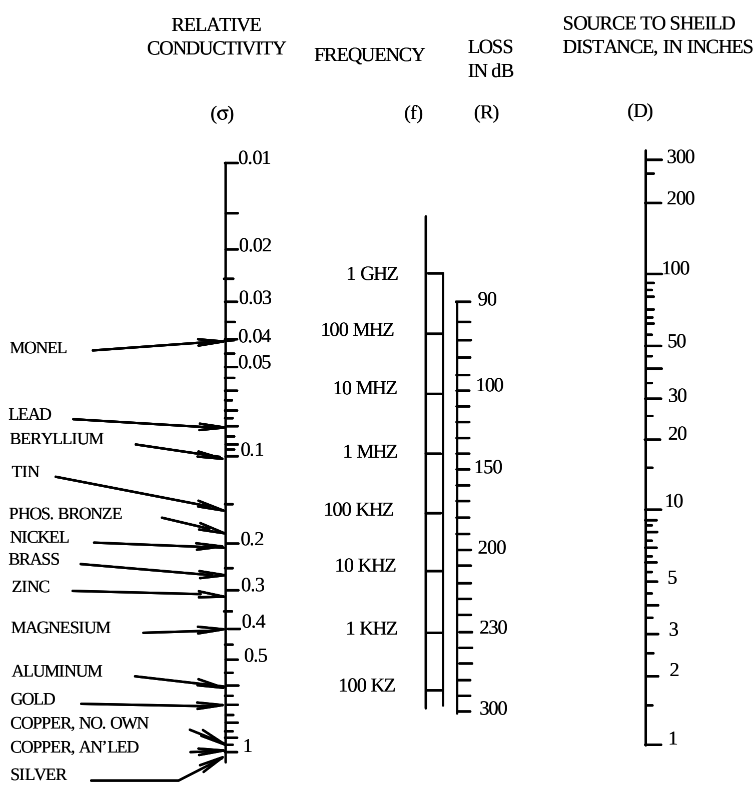


Figure 5-6. Magnetic Field

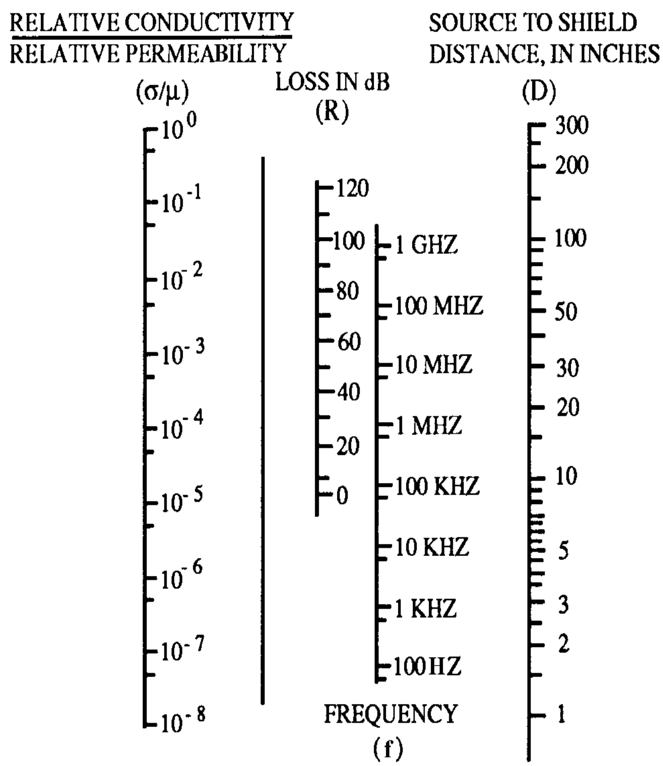


Figure 5-7. Magnetic Field Reflection Loss for Non-Reflection Loss for Magnetic Materials

The impedance of the shield is a complex function of the shield electrical parameters, shield thickness, and frequency of interest; in general:

a. Impedance of a shield is <u>low</u> for materials of high conductivity.

b. Impedance of a shield is <u>high</u> for materials of high permeability.

At radio frequencies of 100 kHz and above, up to 10 GHz or more, the thickness of the shield is generally great enough and the shield hole fraction small enough that the shield impedance can be approximated by that of the solid metallic surface of relative conductivity g and relative permeability μ. This impedance is:

$$2\pi(10^{-7})\sqrt{0.4f\mu/g} \qquad (5\text{-}6)$$

For a non-magnetic shield, $\mu \simeq 1$. For most shield materials (aluminum, copper, silver, etc.), $g \simeq 1$. Thus, even at 10 GHz, the surface impedance is only on the order of 0.04 ohms.

In order that the reflected wave be as large as possible, or that the reflection loss be great, the shielding should have an impedance that is either very much greater than the wave impedance or very much less to maximize the mismatch. In shielding against plane waves (very high impedance), it is more practical to establish a mismatch by using shielding material having a very low impedance than it is to use very high impedance material, and that is why conducting material is always employed.

In the case of high or low impedance sources, it can be seen from equations (5-3) and (5-4) that the reflection loss obtained will be a function of the separation distance between the source and the shield. The equations have been derived on the assumption that these sources are point sources (that is, small compared to the wavelength of concern). The reason the loss is distance-dependent is because we assume that the shield is in the near field $(r < 2D^2/\lambda)$, where the impedance of the wave (E/H) impinging on the shield cannot be defined but must be approximated as a function of r.

In the case of a plane wave source, it can be seen from equation (5-5) that the reflection loss obtained will not be a function of the separation distance between the source and the shield. The reason the loss is distance-independent is because we assume that the shield is in the far field $(r < 2D^2/\lambda)$, where the impedance of the wave (E/H) impinging on the shield is defined as $E = 120\pi H$, which is not a function of r.

The multiple reflection correction term, B, is defined by the equation:

$$B \ (in \ dB) = 20\log\left|1 - 10^{-A/10} W(\cos 0.23A - j\sin 0.23A)\right| \tag{5-7}$$

where

$$W = 4\left[\frac{(1-m^2)^2 - 2m^2 - j2\sqrt{2}m(1-m^2)}{[1+(1+\sqrt{2}m)^2]^2}\right] \tag{5-7a}$$

$$m = 0.776r\sqrt{fg/\mu} \tag{5-7b}$$

$$B = 20 \ \log(1 - e^{-2t/\delta}) \text{ for thin shields} \tag{5-7c}$$

where

$t = thickness \ and \ d = skin \ depth$

t = thickness and d = skin depth

As indicated previously, the multiple reflection correction term may be ignored if the absorption loss exceeds 15 dB. If the absorption loss is less than 15 dB, the correction must be determined.

The multiple reflection correction term is complicated to compute, and depends on material, dimensional, and frequency parameters. Fortunately, in almost all practical cases, W in equation (5-7a) equals unity. The only notable exception is the special case of extremely low-frequency shielding against low-impedance (chiefly magnetic) fields.

Equation (5-7) is combined with the absorption loss and is plotted in figure 5-11 for $W = 1$. Note that the sign of B can be positive or negative, depending on the phase relationships between the reflections. When W is not equal to unity, its value may be obtained using figure 5-12.

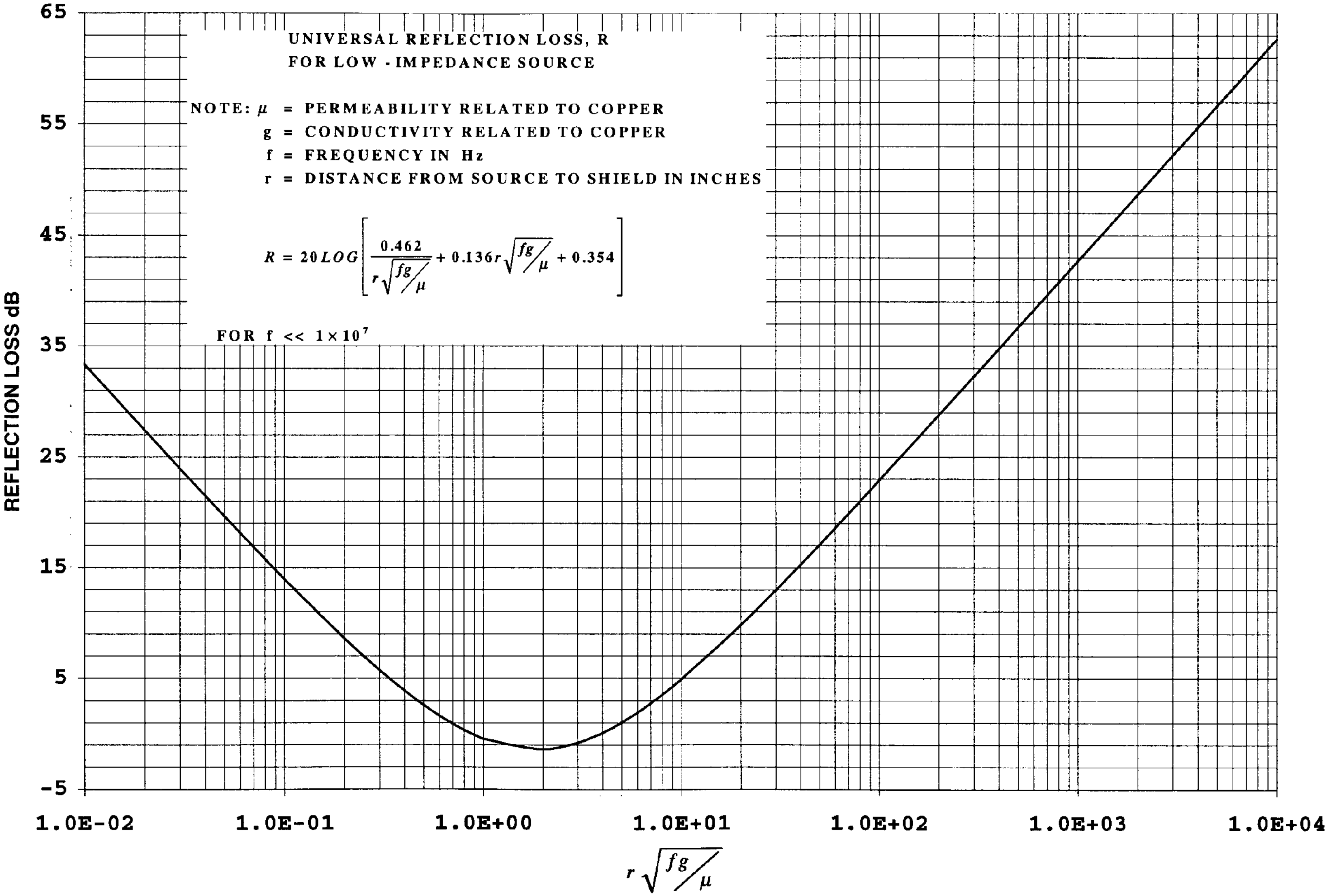


Figure 5-8. Universal Reflection Loss for Low-Impedance Source

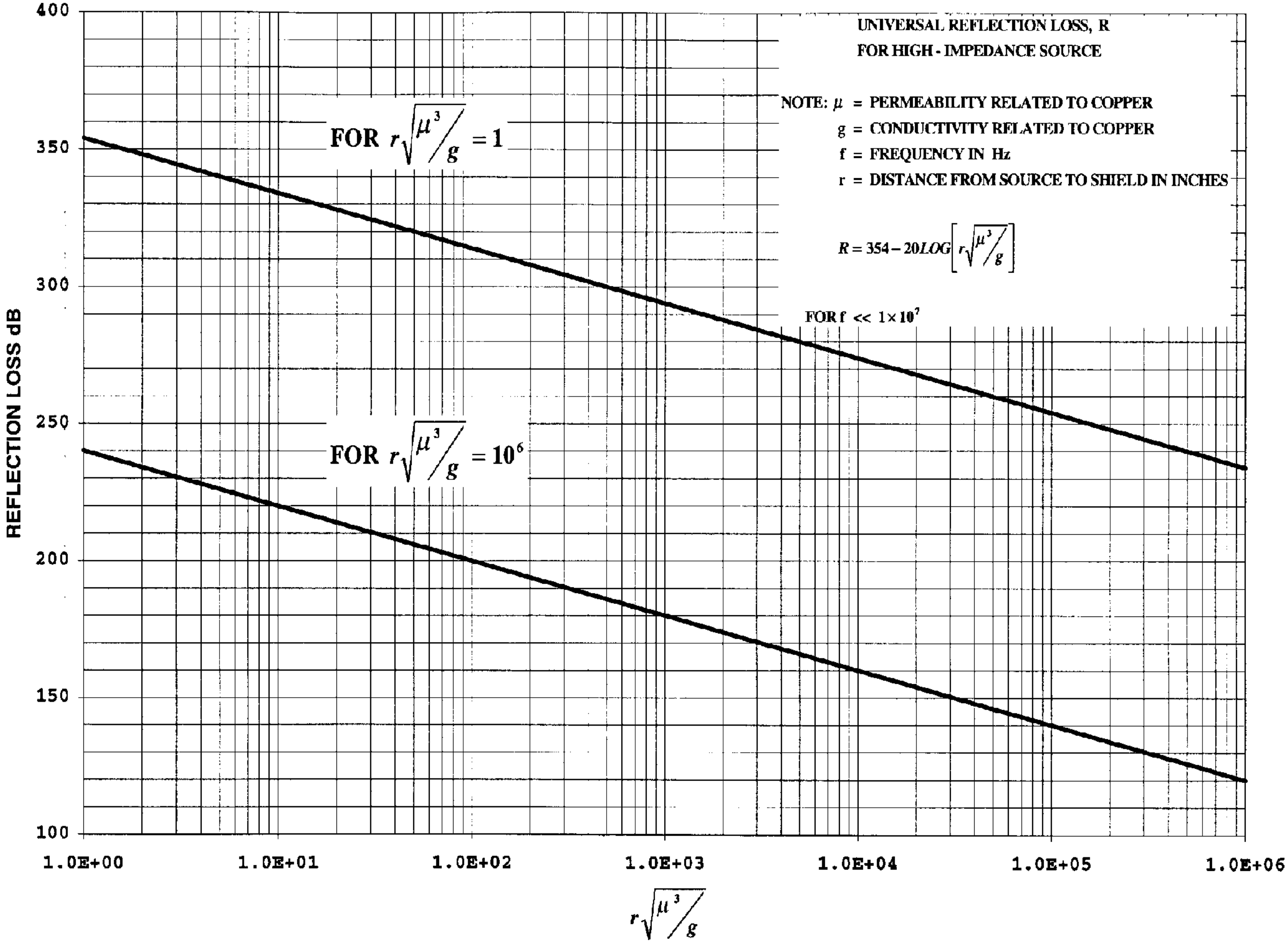


Figure 5-9. Universal Reflection Loss for High-Impedance Source

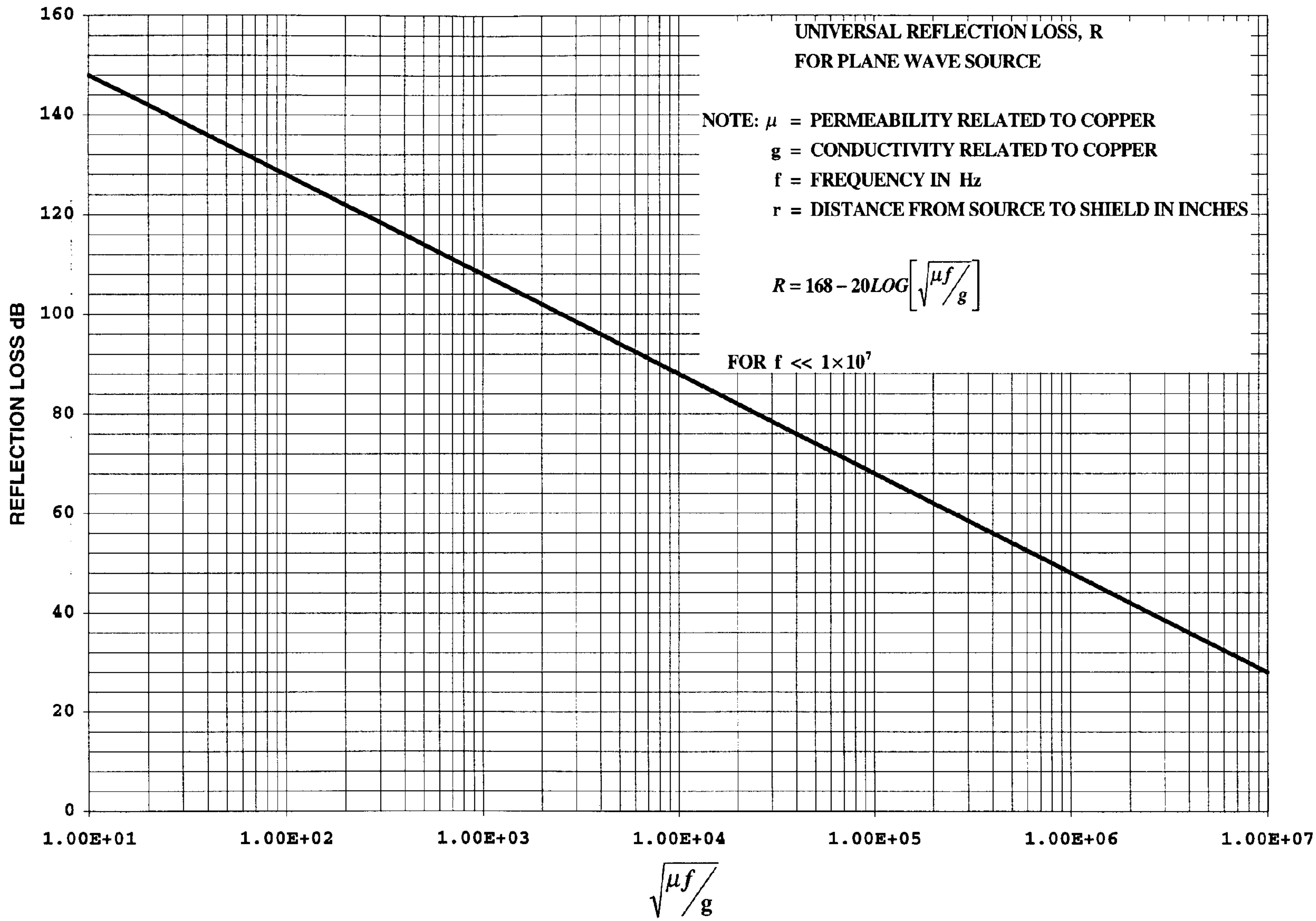


Figure 5-10. Universal Reflection Loss for Plane-Wave Impedance Source

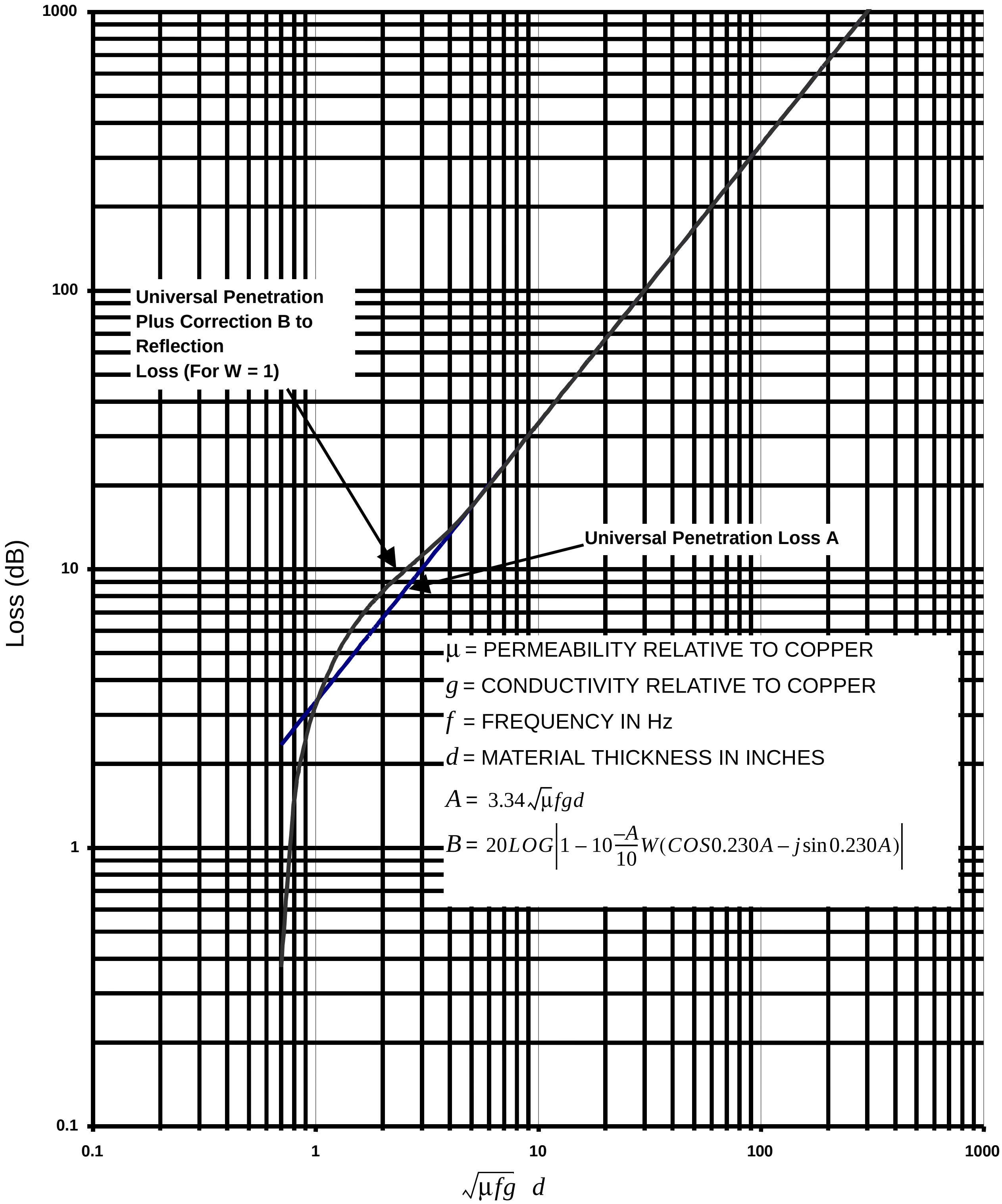


Figure 5-11. Penetration Loss and Multiple Reflection Correction Term when W=1

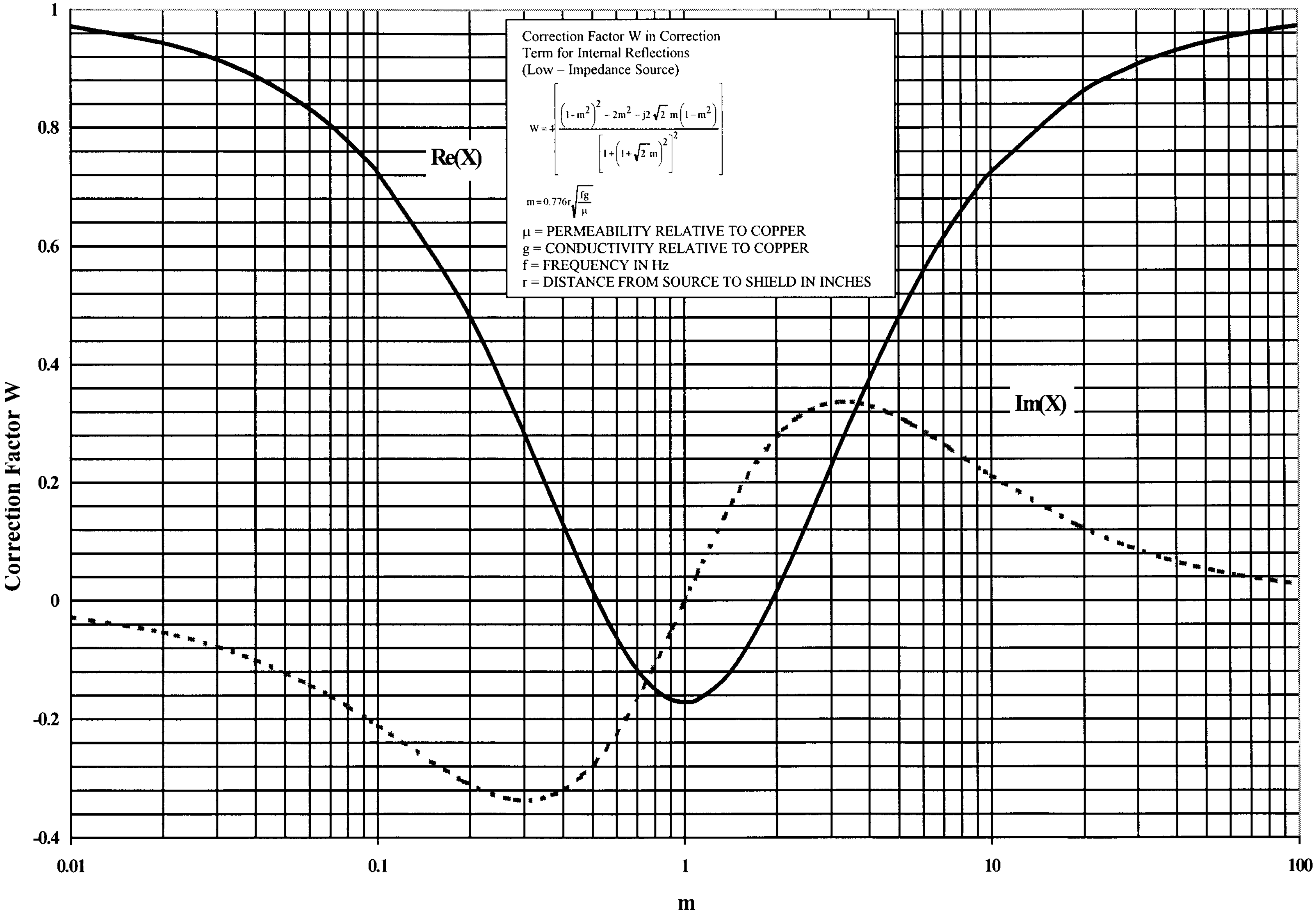


Figure 5-12. Correction Factor in Correction Term for Internal Reflections

Table 5-2 illustrates a typical set of shielding effectiveness calculations for solid sheets of copper and iron. Note the relative magnitudes of the reflection and penetration components. Reflection loss in the electric field cases decreases inversely with frequency.

Representative solid sheet shielding effectiveness measurements are shown in table 5-3. The data encompass a variety of materials subjected to either low or plane-wave impedance energy at frequencies ranging from 100 Hz to 10 MHz. For the materials shown, the shielding effectiveness against high-impedance waves would exceed the plane-wave cases.

5-3.2. COATINGS AND THIN-FILM SHIELDING. Thin shielding has been employed in a variety of ways, ranging from metallized component packaging for protection against RF fields during shipping and storing, to conductive glass, to vacuum-deposited shields for microelectronics applications. Since future ordnance system developments might incorporate thin solid shields, some comments regarding such shields are appropriate.

NOTE

Thin shielding is loosely defined as shield thickness less than one quarter wavelength at the propagation velocity dictated by the material.

Table 5-2. Typical Calculated Values of Shielding Effectiveness of Solid Sheets

Material	Frequency	Type of Field	Metal Thickness (mils)	R Reflection Loss (dB)	A Penetration Loss (dB)	B Correction Term (dB)	Total Shielding Effectiveness S= R + A + B (dB)
Copper	60 Hz	Magnetic	1	22.4	0.026	-22.2	0.23
	60 Hz	Magnetic	10	22.4	0.26	-19.2	3.46
	60 Hz	Magnetic	300	22.4	7.80	+0.32	30.52
	1 kHz	Magnetic	10	34.2	1.06	-10.37	24.89
	10 kHz	Magnetic	10	44.2	3.34	-2.62	44.92
	10 kHz	Electric	10	212.0	3.34	-2.61	212.73
	10 kHz	Plane Waves	10	128.0	3.34	-2.61	128.73
	10 kHz	Magnetic	30	44.2	10.02	+0.58	54.80
	150 kHz	Magnetic	10	56.0	12.9	+0.5	69.4
	150 kHz	Electric	10	176.8	12.9	+0.5	190.2
	150 kHz	Plane Waves	10	117.0	12.9	+0.5	130.4
	1 MHZ	Magnetic	10	64.2	33.4	0	97.6
	1 MHZ	Electric	10	152.0	33.4	0	185.4
Copper	1 MHZ	Plane Waves	10	108.2	33.4	0	141.6

Table 5-2. Typical Calculated Values of Shielding Effectiveness of Solid Sheets (Continued)

Material	Frequency	Type of Field	Metal Thickness (mils)	R Reflection Loss (dB)	A Penetration Loss (dB)	B Correction Term (dB)	Total Shielding Effectiveness S= R + A + B (dB)
Iron	60 Hz	Magnetic	1	-0.9	0.334	+0.95	0.38
	60 Hz	Magnetic	10	-0.9	3.34	+0.78	3.22
	60 Hz	Magnetic	300	-0.9	100.0	0	99.1
	1 kHz	Magnetic	10	0.9	13.70	+0.06	14.66
	10 kHz	Magnetic	10	8.0	43.5	0	51.5
	10 kHz	Electric	10	174.0	43.5	0	217.5
	10 kHz	Plane Waves	10	90.5	43.5	0	134.0
Iron	10 kHz	Magnetic	30	8.0	130.5	0	138.5

Solid material shielding theory is applicable to thin-film shields. For shields much thinner than $\lambda/4$, the absorption loss is very small, but the Multiple Reflection Correction Term, B, is fairly large and negative, thus offsetting a portion of the Single Reflection Loss. The implication of the negative term is that the various reflections have additive phase relationships, and thus reduce the effectiveness of the shield. The shield effectiveness is essentially independent of frequency.

When the shield thickness exceeds $\lambda/4$, the Multiple Reflection Term becomes negligible, and there is no offsetting effect to the other losses. Thus, the material shielding effectiveness increases, and is frequency-dependent.

Table 5-3. Measured Shielding Effectiveness Data for Solid Sheet Materials

Impinging Wave	Material	Thickness (mils)	Nominal Effectiveness							
			0.1 kHz	1 kHz	10 kHz	85 kHz	200 kHz	2 MHz	5 MHz	10 MHz
Plane	Cu	2.5					109	106	112	
Plane	Al	5					107	109	118	
Plane	Stainless Steel	18					97	95	99	
Plane	Steel (U_r =250)	4.5					105	99	101	
Plane	AA-Conetic Foil (U_r -10,000)	3.5					97	130		
Low Impedance	Cu	125 63 31 4.5	8 4	22 11	58 29	97 59 34		120 55		

Table 5-3. Measured Shielding Effectiveness Data for Solid Sheet Materials (Continued)

Impinging Wave	Material	Thickness (mils)	Nominal Effectiveness							
			0.1 kHz	1 kHz	10 kHz	85 kHz	200 kHz	2 MHz	5 MHz	10 MHz
Low Impedance	Al	125 63 31	5 1 1	18 16 10	50 35 24	78				
Low Impedance	Steel (u=242)	63 31	25 4	40 28	80 59	94		92		120
Low Impedance	Brass	31				42				
Low Impedance	Cu-Clad Steel Clad 2 Sides	31				107				
Low Impedance	Cu-Clad Steel Clad 1 Side	94				103				

Table 5-4 provides representative calculations of the shielding effectiveness of thin-film copper for different thicknesses and frequencies, using equations (5-1), (5-2), (5-5), and (5-7). One-quarter wavelength in copper is approximately 32,500 Angstroms (1 Angstrom = 3.94 x 10^{-9} in.) at 1 GHz, and it can be seen that shield effectiveness changes significantly above this thickness.

NOTE

One quarter wavelength = $0.112/(f\mu g)^{1/2}$ in meters.

Table 5-4. Calculated Values of Copper Thin-Film Shielding Effectiveness Against Plane-Wave Energy

Shield Thickness	1050 Å		12500 Å		21960 Å		219600 Å	
Frequency	**1 MHz**	**1 GHz**	**1 MHz**	**1 GHz**	**1 MHz**	**1 GHz**	**1 MHz**	**1 GHz**
Absorption Loss, A	.014	.44	.16	5.2	.29	9.2	2.9	92
Single Reflection Loss, R	109	79	109	79	109	79	109	79

Table 5-4. Calculated Values of Copper Thin-Film Shielding Effectiveness Against Plane-Wave Energy (Continued)

Shield Thickness	1050 Å		12500 Å		21960 Å		219600 Å	
Frequency	**1 MHz**	**1 GHz**	**1 MHz**	**1 GHz**	**1 MHz**	**1 GHz**	**1 MHz**	**1 GHz**
Multiple Reflection Correction Term, B	-47	-17	-26	-.6	-21	.6	-3.5	0
Shielding Effectiveness, S	62	62	83	84	88	90	108	171

Conductive coatings can be applied in a number of ways. Vacuum metallizing involves use of vacuum deposition equipment to apply a metallic film (usually aluminum) onto a surface; thicknesses of about 10^4 Angstroms are typical.

Flame spraying is a somewhat less expensive and time-consuming process than vacuum deposition and typically results in film thicknesses of 2.5×10^5 to 5×10^5 Angstroms for acceptable conductivity. One version consists of two consumable wires that are fed through electrode tips of a spray gun, where a 600-ampere electric current is transferred to them. The arc melts the consumable wires, and an air nozzle behind the wires atomizes the molten metal.

In addition to flame-spraying, conductive coatings can be applied in a number of other ways. These include:

a. Paint consisting of an acrylic-type carrier with silver-metallic or other conductive filler. The paint is sprayed on.

b. Plating of either rigid materials, or films of plastic.

c. Conductive tape or foil, with or without adhesive backing.

Conductive glass (a very thin conducting surface on a clear glass substrate) is often used in applications that require EM shielding properties but must maintain certain visual requirements. A typical shield thickness in this application is 10^4 Angstroms. A more detailed discussion of the applications of conductive glass is presented in paragraph 5-6.5.

5-4. NON-SOLID SHIELDING MATERIALS.

5-4.1. TYPES OF DISCONTINUITIES. An ideal shielded enclosure would be one of seamless construction with no openings or discontinuities. However, personnel, power lines, control cables, and ventilation ducts must have access to any practical enclosure. The design and construction of these discontinuities become very critical in order to incorporate them without appreciably reducing the shielding effectiveness of the enclosure. Some types of discontinuities commonly encountered include:

a. Seams between two metal surfaces, with the surfaces in intimate contact (such as two sheets of material that are riveted or screwed together);

b. Seams or openings between two metal surfaces that may be joined using a metallic gasket;

c. Holes for ventilation, or for exit or entry of wire, cable, light, film, water, meter faces, etc.; and

d. Screened openings.

Design construction considerations associated with such openings will be discussed in the next sections of this design guide.

5-4.2. SHIELDING ANALYSIS. Leakage through openings in metal shields has been studied using transmission line theory. Based on these studies, the Shielding Effectiveness, S, of non-solid shielding materials has been defined as:

$$S = A_a + R_a + B_a + K_1 + K_2 + K_3 \tag{5-8}$$

where

A_a = Absorption (Penetration) loss introduced by a particular discontinuity, in dB

R_a = Aperture Single Reflection Loss, in dB

B_a = Multiple Reflection Correction Term, in dB

K_1 = A correction term to take into account the number of like discontinuities

K_2 = A low-frequency correction term to take into account skin depth

K_3 = A correction term to take into account coupling between adjacent holes.

The first three terms in equation (5-8) generally correspond to the three terms of equation (5-1), while the last three terms encompass other factors that need not be considered for solid sheets. The analysis is most appropriate for single discontinuities, or for identical and uniformly spaced apertures (such as screening or perforated sheets), but can be extended to somewhat more complex configurations as well.

The absorption loss term, A_a, is obtained by assuming that the impinging wave is well below its cutoff frequency.

NOTE

The cut-off frequency (F_c) is that frequency below which the propagation constant of the aperture is real. It is defined by $f_c = C/\lambda_c$, where c is the velocity of light and λ_c is the cut-off wavelength, in the same units. The cut-off wavelength is equal to twice the maximum dimension of a rectangular aperture, or 3.142 times the radius of a circular aperture.

Under these circumstances:

$$A_a = 27.3\ d/w \quad \text{(for rectangular aperture)} \tag{5-9a}$$

$$A_a = 32\ d/D \quad \text{(for circular aperture)} \tag{5-9b}$$

where

d = depth of aperture, in inches

w = width of aperture perpendicular to the E-field vector, in inches

D = aperture diameter, in inches.

These equations are plotted in figure 5-13A.

The multiplying factor to be applied to equations (5-9a and 5-9b) for frequencies f not well below cut-off is $[1-(f/f_c)^2]^{1/2}$.

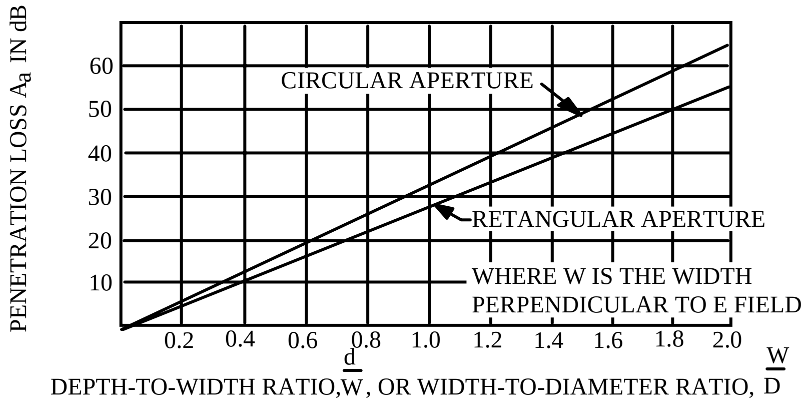


Figure 5-13A. Aperture Shielding Absorption Loss

The aperture single reflection loss, R_a, is wave-impedance dependent and is also a function of aperture shape. It can be expressed as:

$$R_a = 20\log\left[\frac{1+|K|^2}{4|K|}\right] \quad (5\text{-}10)$$

where

$K = w/\pi r$, for low impedance fields and rectangular apertures

$K = D/3.682\, r$, for low impedance fields and circular apertures

$K = jfw(1.7 \times 10^{-4})$, for plane-wave fields and rectangular apertures

$K = jfd(1.4 \times 10^{-4})$, for plane-wave fields and circular apertures

f = frequency, in MHz

r = source to shield distance, in inches

$j = \sqrt{-1}$

Graphs of R_a for the low-impedance case are shown in figure 5-13B.

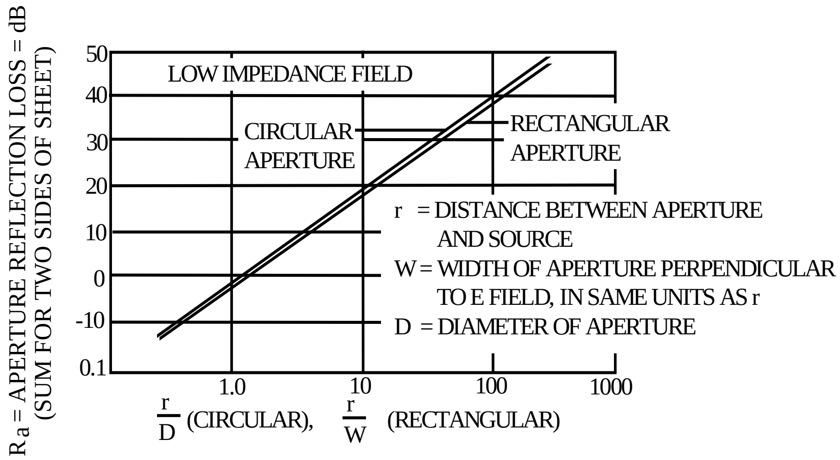


Figure 5-13B. Aperture Shielding Reflection Loss

The <u>multiple reflection correction term</u>, B_a, is defined by the equation:

$$B_a = 20\log\left[1 - \frac{(K-1)^2}{(K+1)^2}10^{-Aa/10}\right] \qquad (5\text{-}11)$$

and is applicable only when A_a is less than 15 dB.

The <u>number of discontinuities correction term</u>, $K1$, is employed when the source is located a large distance from the shield, in comparison with the aperture spacing in the shield. It is represented by:

$$K_1 = -10\log(an) \qquad (5\text{-}12)$$

where

a = area of each hole, in square inches

n = number of holes per square inch

The <u>skin depth correction term</u>, K_2, recognizes that, at low frequencies, when the skin depth becomes comparable to the screening wire diameter or dimension between apertures, a reduction in shielding effectiveness can occur. An empirical relationship for this effect was developed:

$$K_2 = -20\log(1 + 35/p^{2.3}) \tag{5-13}$$

where

p = the ratio of wire diameter to skin depth, for screening, or

p = the ratio of conductor width between holes to skin depth, for perforated sheets

For convenience, the equation for skin depth in copper, in inches, is

$$SD_{CU} = 2.6 \times 10^{-3}\sqrt{f} \tag{5-14}$$

The adjacent hole coupling correction term, K_3, is the result of noting that shielding efficiency is higher than expected when apertures in a shield are closely spaced and the depths of the openings are small compared to the aperture width. This is interpreted as the result of coupling between adjacent holes, and becomes important for small openings. The equation for computing K_3 is:

$$K_3 = -20\log[\coth(A_a/8.686)] \tag{5-15}$$

The elements of the non-solid shielding effectiveness equation just discussed are summarized in table 5-5 for convenience. Calculations in accordance with these equations have resulted in fairly good agreement with measurements, so long as the aperture pattern is uniform. Table 5-6 also provides similar comparative data.

Table 5-5. Terms for Aperture Shielding

Symbol	Item	Aperture		Comments
		Rectangular	Circular	
A_a	Penetration Loss	$27.3\frac{d}{w}$	$32\frac{d}{D}$	d =Depth of Aperture, inches w =Width of Rectangular Aperture, Perpendicular to E field, inches D =Dia. of Circular Aperture, inches

Table 5-5. Terms for Aperture Shielding (Continued)

<table>
<tr><th rowspan="2">Symbol</th><th rowspan="2">Item</th><th colspan="2">Aperture</th><th rowspan="2">Comments</th></tr>
<tr><th>Rectangular</th><th>Circular</th></tr>
<tr><td rowspan="3">R_a</td><td rowspan="3">Aperture Reflection Loss (sum for 2 sides of sheet)</td><td>$20\log_{10} 0.785\frac{r}{w}$</td><td>$20\log_{10} 0.925\frac{r}{D}$</td><td><< 377Ω</td></tr>
<tr><td>$20\log_{10}\frac{1+(1.7x10^{-4}Fw)^2}{4(1.7x10^{-4}fw)}$</td><td>$20\log_{10}\frac{1+(1.47x10^{-4}fw)^2}{4(1.47x10^{-4}fw)}$</td><td>377Ω</td></tr>
<tr><td>--------------------</td><td>--------------------</td><td>>>377Ω</td></tr>
<tr><td>B_a</td><td>Correction Term Due to Successive Reflections When A_a <15db</td><td colspan="2">$20\log_{10}\left(1-10^{\frac{Aa}{10}}\right)$</td><td>$\frac{w}{\pi r} 1 \frac{D}{3.68r} \ll 1$</td></tr>
<tr><td>K_1</td><td>Loss Term Due to Number of Openings Per Unit Square</td><td colspan="2">$-10\log_{10} an,\ r \gg \omega, D$

$zero,\ r \ll \omega, D$</td><td>a=Area of Single Aperture, inches^2 n=No. of Holes per sq. inch</td></tr>
<tr><td>K_2</td><td>Correction Term for Penetration of Conductor at Low Frequencies</td><td colspan="2">$-20\log_{10}(1+35p^{-2.3})$</td><td><u>Wire Dia.</u> Skin Depth, Screening

P=Conductor Width Between Holes/Skin Depth Perforated Sheets</td></tr>
<tr><td>K_3</td><td>Correction for Closely Spaced Shallow Holes</td><td colspan="2">$20\log_{10}\left(\coth\frac{Aa}{8.686}\right)$</td><td></td></tr>
</table>

Table 5-6. Comparison of Measurements and Calculations of Screening Material Shielding Effectiveness

Screen Type*	Test Type	Frequency (MHz)	Mean Values (dB)	Calc. Values (dB)
No. 22 15 mil	Magnetic Field	0.085 1.0 10.0	31 43 43	28 45 49
No. 22 15 mil	Plane Waves	0.2 1.0 5.0 100.0	118 106 100 80	124 110 95 70
No. 22 15 mil	Electric Field	0.01 to 60	65	65 **
No. 22 20 mil	Electric Field	0.01 to 60	50	53 **

* All screens made of copper.
** These values assume a wave impedance equal to that of a 30-inch square waveguide.

Representative non-solid sheet shielding effectiveness measurements are shown in tables 5-7 and 5-8. The two tables provide data on a variety of material forms, including meshes, perforated sheets, and cellular structures against low-impedance, high-impedance, and plane waves.

5-4.3. COMPOSITE MATERIALS. In recent years, composite materials have been developed to the point that their strength and weight characteristics make them very attractive as substitutes for metals in various aircraft and ordnance applications. Composites are a family of polymeric or metal-reinforced materials having organic or inorganic fibers or filaments, woven fabric layering, and epoxy bonding. Boron, graphite, Kevlar 49, and glass fibers are employed, with boron and graphite most frequently used. Recent developments include nickel plating graphite fibers, aluminum coating glass or graphite fibers, and graphite coating stainless steel fibers.

Table 5-7. Effectiveness of Non-Solid Shielding Materials Against Low-Impedance and Plane Waves

<table>
<tr><th rowspan="2">IMPINGING WAVE</th><th colspan="2">FORM</th><th rowspan="2">MATERIAL</th><th rowspan="2">THICKNESS (mils)</th><th colspan="6">NOMINAL EFFECTIVENESS (db)</th></tr>
<tr><th>GENERAL</th><th>DETAIL</th><th>0.1 kHz</th><th>1 kHz</th><th>10 kHz</th><th>85 kHz</th><th>1 MHz</th><th>10 MHz</th></tr>
<tr><td rowspan="4">Low Impedance</td><td rowspan="4">Mesh (Screening)</td><td>2 layers 1 inch apart</td><td>Cu (oxidized)</td><td></td><td>2</td><td>6</td><td>18</td><td></td><td></td><td></td></tr>
<tr><td>No. 22</td><td>Cu</td><td></td><td></td><td></td><td></td><td>31</td><td>43</td><td>43</td></tr>
<tr><td>No. 16</td><td>Bronze</td><td></td><td></td><td></td><td></td><td>18</td><td></td><td></td></tr>
<tr><td>No. 4</td><td>Galvanized Steel</td><td></td><td></td><td></td><td></td><td>10</td><td>17</td><td>21</td></tr>
<tr><td rowspan="5">Plane</td><td rowspan="2">Perforated Sheet</td><td rowspan="2"><5 mil dia., 225 sq. inch</td><td rowspan="2">Al</td><td rowspan="2">20</td><td colspan="3">3040 MHz</td><td colspan="3">9380</td></tr>
<tr><td colspan="3">60</td><td colspan="3">62</td></tr>
<tr><td rowspan="3">Mesh (Screening)</td><td>No. 16</td><td>Al</td><td>dia=13</td><td colspan="3">34</td><td colspan="3">36</td></tr>
<tr><td rowspan="2">No. 22</td><td rowspan="2">Cu</td><td rowspan="2">dia=15</td><td>200 kHz</td><td colspan="2">1 MHz</td><td colspan="2">5 MHz</td><td>100 MHz</td></tr>
<tr><td>118</td><td colspan="2">106</td><td colspan="2">100</td><td>80</td></tr>
</table>

Table 5-8. Effectiveness of Non-Solid Shielding Materials against High-Impedance Waves

<table>
<tr><th colspan="2">FORM</th><th rowspan="2">MATERIAL</th><th rowspan="2">THICKNESS (mils)</th><th>NOMINAL EFFECTIVENESS (db)</th><th rowspan="2">OPEN AREA %</th><th colspan="2">AIR FLOW STATIC PRESSURE (inches of water)</th></tr>
<tr><th>GENERAL</th><th>DETAILED</th><th>14 kHz to 1000 MHz</th><th>200 cu ft/min</th><th>400 cu ft/min</th></tr>
<tr><td>Hexcell</td><td>1/4-inch cell, 1 inch thick</td><td>Al</td><td>3</td><td>>90</td><td></td><td>0.06</td><td>.26</td></tr>
<tr><td rowspan="3">TV Shadow Masks (Photo-Etched)</td><td>9-mil holes 28-mil centers</td><td rowspan="2">95%Cu 5%Ni</td><td rowspan="2">7</td><td rowspan="2">>90</td><td>12</td><td>>2</td><td></td></tr>
<tr><td></td><td>50</td><td>0.2</td><td>0.4</td></tr>
<tr><td></td><td>100%Ni</td><td>3</td><td>>90</td><td>50</td><td>0.2</td><td>0.5</td></tr>
</table>

Table 5-8. Effectiveness of Non-Solid Shielding Materials against High-Impedance Waves (Continued)

FORM		MATERIAL	THICKNESS (mils)	NOMINAL EFFECTIVENESS (db)	OPEN AREA %	AIR FLOW STATIC PRESSURE (inches of water)	
GENERAL	DETAILED			14 kHz to 1000 MHz		200 cu ft/min	400 cu ft/min
	40 count		7	>90	36	0.4	1.7
	25 count	Cu-Ni	5	78	49	0.2	0.5
	40 count				57	0.2	0.5
Lektromesh	25 count	CU	3	78	56	0.2	0.4
	1/8-inch dia. 3/16-inch centers	Steel	60	58		0.27	>0.6
	1/4-inch dia. 5/16-inch centers		60	48	46		
Perforated Sheet	7/16-inch dia. 5/8-inch centers	Al	37	35	45		
	No. 16 16 x 16/sq. in.	Al	20 (dia)	55	36		
	No. 22			65(14kHz to 60 MHz)			
	No. 12	Cu	20 (dia)	50	50		
	No. 16	Bronze		45(14 kHz to 60 MHz)			
	No. 10	Monel	18 (dia)	40			
	No. 4			35 28(14 kHz to 40 MHz)	76		
Mesh (Screening)	No. 2	Galvanized Steel	30 (dia)	24	88		

Boron filaments are produced by vapor deposition on a tungsten wire resulting in a very stiff, high strength, low density fiber. The boron composite is very limited in ability to transport electrical current. This fact makes the boron composite susceptible to high currents, such as lightning, and to charge-accumulated high potentials, such as static electricity.

Graphite filaments are made by pyrolysis of polyacrylonitrile, or decomposition of a material similar to rayon. This produces a strong, stiff, low-cost filament. The graphite material is more conductive than boron and, therefore, somewhat less susceptible to electrical phenomena. In addition, graphite possesses basic EM characteristics which, although reduced from metals, are improvements over other composites. The extent to which composites alter the electrical characteristics of the structure in which they are used in weapon systems clearly depends on the amount and location of the material.

The shielding effectiveness of composites is much lower than that of conductive metals. Essentially, the composite provides little protection from flux linkage of lightning energy to internal subsystems and equipment. In addition, little attenuation is offered to other radiated environmental signals. The more limited shielding provided by the airframe transfers the shielding burden to internal subsystems of the aircraft or weapon.

The applications of composite materials in ordnance will continue to have a significant impact on all aspects of electrical system grounding. Since composites are not capable of supporting significant current flow, they cannot serve as a system ground-plane. Alternative approaches, such as use of wired electrical power returns and ground buses, become necessary.

5-5. CABLES AND CONNECTORS.

5-5.1. CABLE SHIELDING. There are several methods for shielding cables. These include (1) braid, (2) flexible conduit, (3) rigid conduit, (4) spirally wound shields of high permeability materials, and (5) encasing the conductor in a ferrite material (filter-line cable).

Braid, which constitutes woven or perforated material, is used for cable shielding in applications where the shield cannot be made of solid material. Advantages are ease of handling in cable make-up and lightness in weight. However, it must be remembered that for radiated fields, the shielding effectiveness of woven or braided materials decreases with increasing frequency and the shielding effectiveness increases with the density of the weave. The percentage coverage by a braided shield has been a critical parameter in the design of ordnance system cables.

Conduit, either solid or flexible, may also be used to shield ordnance system cables and wiring from the RF environment. The shielding effectiveness of solid conduit is the same, for RF purposes, as that of a solid sheet of the same thickness. Shielding degradation of conduit is usually not the result of insufficient shielding properties of the conduit material, but rather the result of discontinuities in the conduit. These discontinuities usually result from poor splicing techniques, poorly bonded interconnects, or improper termination of the shield at connectors or plugs. Physical protection of ordnance cables can be provided by linked armor or flexible, corrugated conduit. These cable conduits do provide some low frequency shielding, but at higher frequencies, serious degradation occurs because the openings between individual links can take on slot-antenna characteristics which can couple EME directly onto the interior cables. If linked armor conduit is required, all internal wiring should be individually shielded.

The principal types of shielded cables that are available include shielded single wire, shielded multiconductor, shielded twisted pair, and coaxial. Cables are also available in both single and multiple shields in many different forms and with a variety of physical characteristics.

One physical configuration is the filter line cable. Similar to annealed metal tape, the shielding material has a high permeability; in this case, however, the material is a ferrite. The ferrite encases the conductor and takes advantage of the loss versus frequency characteristics of this material.

Data on the shielding effectiveness of cables are not easily measured. This is because a large number of parameters (some of them external to the cable itself) dictate the particular performance of a cable, including termination impedances, impinging signal direction and impedance, cable length, frequency, the particular connectors employed, flexing requirements, and others. However, at frequencies where diffusion effects are negligible, the shielding effectiveness can be determined in a way that is invariant with changes in cable length or experimental conditions. The triaxial test method is used to obtain a dB attenuation versus frequency. Then, the Surface Transfer Impedance (STI) and Geometric Transfer Ratio (GTR) are determined in order to accommodate a best fit between a calculated spectrum and the measured spectrum. This enables predictions of the shield leakage of a cable at any frequency, for any length, and under any definable set of test conditions.

The general characteristics of four classes of shielded cables are identified in table 5-9. The classes include rigid and flexible conduit, foil-wrapped cable, and braided shielded cable. As indicated previously, shielding effectiveness in most cabling applications is dependent on the percentage coverage of the cable provided by the shield.

For example, MIL-C-7078/35 Notice 1 calls for the percent shield coverage (the ratio of metal to total possible shield surface) to be at least 85 percent. It also imposes a requirement on the angle of the braid wires relative to a plane normal to the cable axis (a parameter that influences percentage shield coverage under cable flexing conditions) of between 10 and 40 degrees.

Table 5-9. Comparison of Shielded Cables

	COPPER BRAID	FOIL B	SOLID CONDUIT	FLEXIBLE CONDUIT
Shield Effectiveness (audio frequency)	Good	Excellent*	Excellent	Good
Shield Effectiveness (radio frequency)	Good	Excellent*	Excellent	Poor
Normal % of Coverage	60-95%	100%	100%	90-97%
Fatigue Life	Good	Fair	Poor	Fair
Tensile Strength	Excellent	Poor	Excellent	Fair

* May lose its effectiveness when flexed.

For ordnance applications, tests made on particular systems indicate it is necessary to impose a shield coverage requirement of 94 percent on braided cable. The shield braid angle and percent braid coverage are determined in accordance with MIL-C-7078C, as follows:

$$\text{Tan } a = 2\pi(D + 2d)P/C \tag{5-16}$$

$$K = 100(2F - F^2) \tag{5-17}$$

where

a = Shield braid angle (angle of braid with axis of cable)

K = Percent coverage

d = Shield strand diameter in inches

C = Number of carriers

D = Diameter of cable under braid, in inches

$F = NPd/\text{Sin } a$

N = Number of strands per carrier

P = Picks per inch of cable length

Note: For 2-conductor cable only

$$D = \frac{(\pi + 2)b}{\pi} \tag{5-18}$$

where

b = diameter of basic wire

Some of these parameters are defined in figure 5-14.

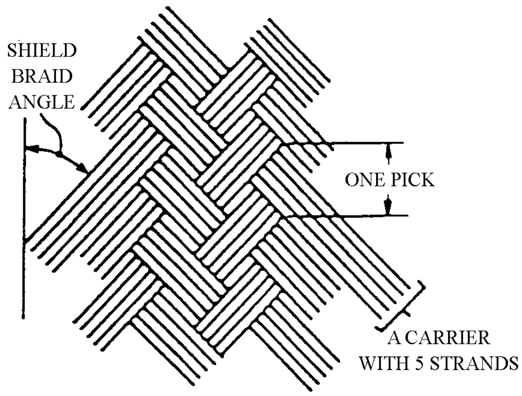


Figure 5-14. Definitions of Cable Shield Parameters

Much of the previous discussion relates to the shielding of cables against plane wave or high-impedance fields. For shielding against magnetic fields, the use of annealed high-permeability metal strips wrapped around the cable has already been indicated. Multiple layers of counterspiral-wound nickel-iron or silicon iron alloys, or low carbon steel have proven effective under these circumstances. High permeability tape is available with or without adhesive backing.

Also, combination high permeability, high conductivity tape is available to provide both electric and magnetic shielding.

5-5.2. CABLE SHIELD TERMINATIONS AND CONNECTORS. If the effectiveness of a shield is to be maintained, the cable shield must be properly terminated. In otherwise adequately shielded systems, RF currents that are conducted along shields can be coupled to the system wiring from the point of an improper cable termination. This is a particularly important consideration in the case of cables exposed to high power RF fields. In fact, one school of thought is that HERO is affected mostly by system cable effects, rather than leakage through cracks, seams, etc. In one study conducted, it was found that a shielded conductor carrying an RF current induces the negative of that current on the inside surface of the shield. When this return current is confined to the inside surface, the shielding is most effective. When this return current is not confined to the inside surface, some of it flows down the outside surface reducing shielding effectiveness.

Thus, in a properly terminated shield, the entire periphery of the shield is grounded to a low impedance reference, minimizing any RF potentials at the surface of the termination. Solder is undesirable in terminating RF coaxial cables because too much solder increases the center conductor diameter, thus increasing shunt capacitance, and too little solder increases the current path, thus increasing series inductance. Specification MIL-E-45782B recommends against use of soldering to terminate shields because of the danger of damaging conductor insulation, and suggests a variety of termination methods, all involving crimping operations. A frequently used method of shield termination is illustrated in figure 5-15. In this arrangement, the cable shield is flared so that it extends over the rear portion of the sleeve, and the crimp ring is slid into place over the sleeve. A crimping tool is then used to crimp the crimp ring onto the sleeve.

An alternative to crimping is shown in figure 5-16. The shield is placed through the ground ring and flared over and around the ring, and may be secured to the ring with a spot tie (see detail in figure 5-16). The ground ring is then slid into the rear of the sleeve, which has a tapered base. Tightening the cable clamp onto the end of the sleeve assures positive 360 degrees grounding of the shield, and provides a strain relief for the cable. The use of silver epoxy or other synthetic conducting material has been found to be unacceptable for shield bonding because of lack of mechanical strength necessary for this application. A variation on this termination method includes environmental seals in the backshell construction. This is of particular value for ordnance that will operate in the corrosive marine environment aboard ship.

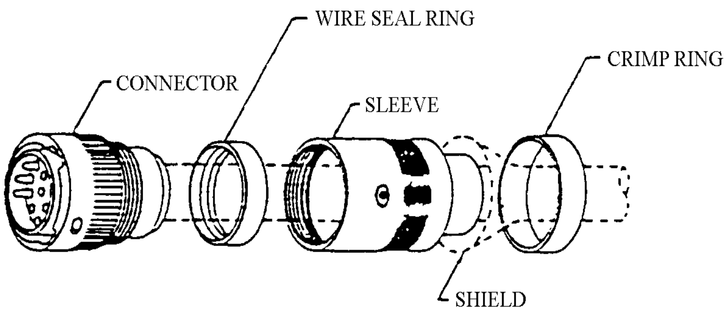


Figure 5-15. Shield Termination Using Crimping

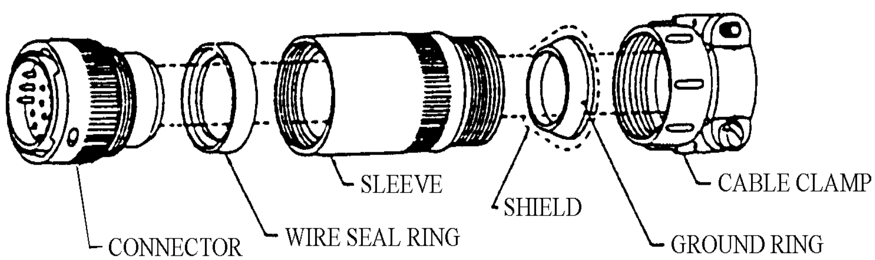


Figure 5-16. Shield Termination Using Threaded Assembly

Since ordnance must operate in high EME's, both shields and solid cylindrical members have been terminated using the Magnaforming process. Magnaforming is a metal-forming technique that is used to shrink metal tubes and similar shapes around other forms such as collars, sleeves, rods, etc.

When maintaining the shielding integrity of the connector pair (i.e., two interconnecting connectors), a good method to employ (see figure 5-17) is to place spring contacts inside one portion of one connector so that positive contact is made along the circumference of the mating parts. These contacts are extended so that the shell of the connector mates before the pins make contact on assembly of the connector and breaks after the pins on disassembly. A connector which meets these requirements is available under MIL-C-27599 and is the preferred type to be used in RF-proof designs.

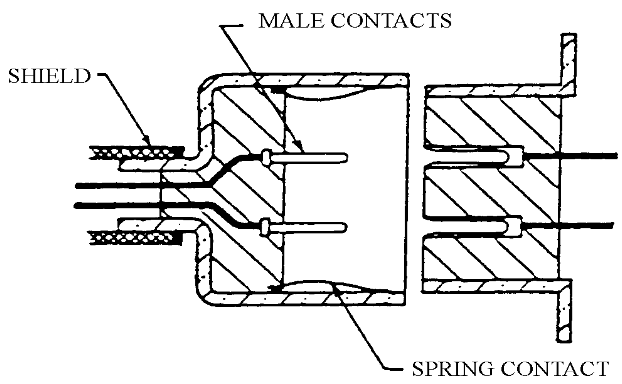


Figure 5-17. RF-Proof Connector

Figure 5-18 illustrates the type of connector that should be used when a shielded cable assembly contains individual shielded wires. The practice of pigtailing these shields and connecting them to one of the pins is not recommended. The individual shields should be connected to coaxial pins specifically adapted for this purpose, with the shields of the mating surfaces making contact before the pins.

Two types of connectors with special features that improve shell-to-shell bonding in a high vibration environment are the Bendix "Tri-Start" series and the G & H Technology "Breech-Lok" series. Both are "scoop-proof" connectors that employ metric threads at the rear for attaching accessories. Scoop-proofing prevents shorts and pin damage, and metric threads provide more thread surface and thus a better bond between shells and accessories.

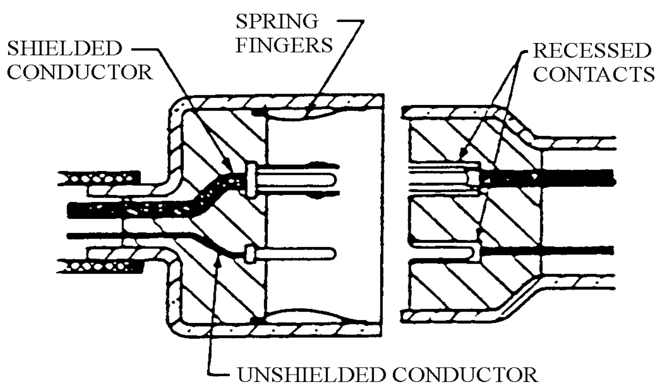


Figure 5-18. Connector for Shield within a Shield

The Tri-Start (Bendix Corporation) connectors are Series III connectors as defined in MIL-C-38999G and DOD-C-38999. They incorporate Series I high-density inserts and contacts, and use a triple start Acme thread (one full turn to lock) with an anti-decoupling feature for plug-to-receptacle engagement. They also include metallic spring fingers for grounding. They were developed to meet rigid environmental and shock vibration requirements, but also provide improved EMI protection over the bayonette type Series I and II connectors in a high-vibration environment.

The Breech-Lock (G&H Technology, Inc.) connectors are Series IV connectors as defined in MIL-C-38999G and DOD-C-38999. They incorporate Series I inserts and contacts, but use a solid-metal locking system (detent and lock in one-half turn) and internal drive threads for plug-to-receptacle engagement. They also include metallic spring fingers for grounding. They too were developed to meet rigid environmental and shock and vibration requirements, but also provide improved EMI protection over Series I and II connectors in a high-vibration environment.

Another connector technology is the “filter connector.” Although they were developed in the 1960's, filter connectors have only been available as production items since about 1980. Filter connectors are used to supplement shielded cable and shielding connectors when they cannot reduce incoming or outgoing interference to acceptable levels.

5-5.3. CONNECTOR TECHNOLOGY. Some connector technologies worth noting include conduit compatible backshells, EMI/RFI shrink boot adapters, and conductive-coated composite connectors.

The Breeze Illinois, Inc. BI-SHELL 10/11 backshell can accommodate flexible metallic and non-metallic conduit to provide a hermetic/EMI seal. Both conduit and braided shield are inserted into the end of the backshell and terminated by brazing, or soldering the seam; the termination can also be accomplished by magnaforming. If environmental protection is required, an elastomer jacket can cover the underlying braided shield and flexible conduit slipping over the backshell collar. This jacket would then be terminated by bonding.

Glenair Inc. manufactures EMI/RFI shrink boot adapters which feature a unique termination scheme. Shrink boots fit over the junction of the backshell and cable and "shrink" when heated to provide mechanical strain relief and some environmental protection. The unique termination scheme for these particular adapters is that the backshell collar and termination nut have sine wave threads which lock the braided shield between them when the shield is pulled over the collar threads and the termination nut is tightened (see figure 5-19).

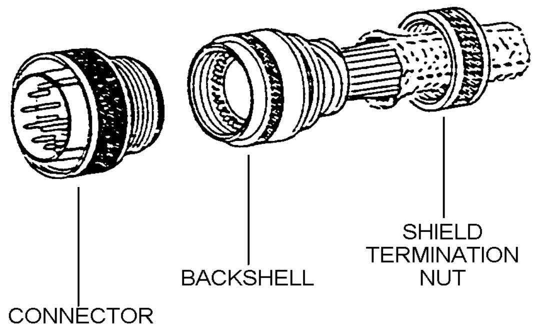


Figure 5-19. EMI/RFI Shrink Boot Adapter

Deutsch manufactures the DG123 conductive-coated composite connector. This corrosion resistant connector passes MIL-C-38999, class W series III shielding effectiveness tests, and has a shell design based on the insert arrangement patterns of MIL-C-38999 and MIL-C-81511 connectors. This connector is capable of 1500 mating cycles without degradation and has features contributing to extended life cycle such as molded dielectric fingers in its inserts (for retention), and keys (for keyways) which are molded into the shell.

5-5.4. FIBER OPTICS FOR ORDNANCE. Fiber optics technology is being incorporated into advanced weapon systems and in associated test equipment. It makes possible optical links between subsystems, by replacing hard-wired interfaces with glass fibers; and transmitting voice, command and control, video or data signals by light rather than by electricity. Infrared lasers or light-emitting diodes serve as signal drivers. From the standpoint of the HERO problem, fiber-optics cables are insensitive to EM field environments (although that is not necessarily true of the line termination devices) and therefore do not have to be shielded. They provide electrical isolation between source and load, eliminate potential problems of ground loop coupling, and can eliminate EME coupling between test sites and missile or platform/ launcher and missile. Additionally, they are totally safe for use in a fuel vapor or other explosive environment.

The electromagnetic compatibility (EMC) features must be weighed against other characteristics associated with fiber-optics subsystems. These include the more complex and more costly line termination equipment, and the energy loss characteristics of the fibers themselves as compared with copper wire. The fibers can be clad and bundled in arrangements very similar to hard-wired cable.

5-6. OTHER DESIGN TECHNIQUES TO MAINTAIN SHIELDING EFFECTIVENESS.

5-6.1. SEAMS WITHOUT GASKETS. Seams or openings in enclosures or compartment walls that are properly connected together will provide good shielding effectiveness by creating a low impedance path to RF currents flowing across the seam. When high quality shielding characteristics are to be maintained, permanent mating surface of metallic members within an enclosure should be bonded together by welding, brazing, sweating, swaging, or other metal flow processes.

Several configurations which form seams between two metallic members within a weapons system are shown in figure 5-20. The preferred seam is a continuous weld around the periphery of the mating surfaces. The type of weld is not critical, provided the weld is continuous. Spot welding can also be used, provided care is exercised to prevent gaps in the mating surfaces between the spot welds. A good guide line is having the spacing of the spot welds to be not greater than the overlap of the surfaces being joined.

An acceptable alternative technique is the crimp seam pictured in figure 5-21. In a crimp seam, all non-conductive materials must be removed from the mating surfaces before the surfaces are crimped. The crimping must be performed under sufficient pressure to ensure positive contact between all mating surfaces.

To ensure adequate and proper shielding effectiveness of seams without gaskets, the following recommendations should be observed:

a. All mating surfaces must be cleaned before bonding.

b. All protective coatings having a conductivity less than that of the metals being bonded must be removed from the contact areas of the two mating surfaces before the bond connection is made.

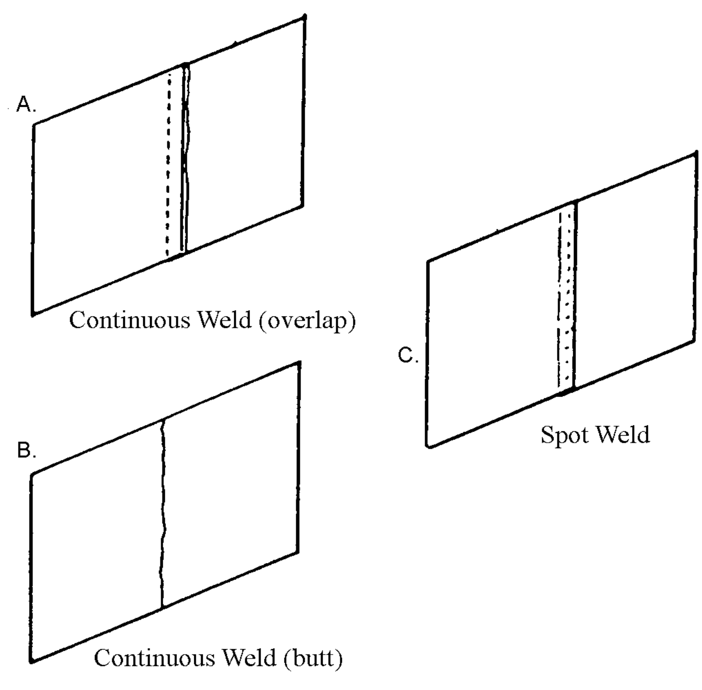


Figure 5-20. Panel Seam Configurations

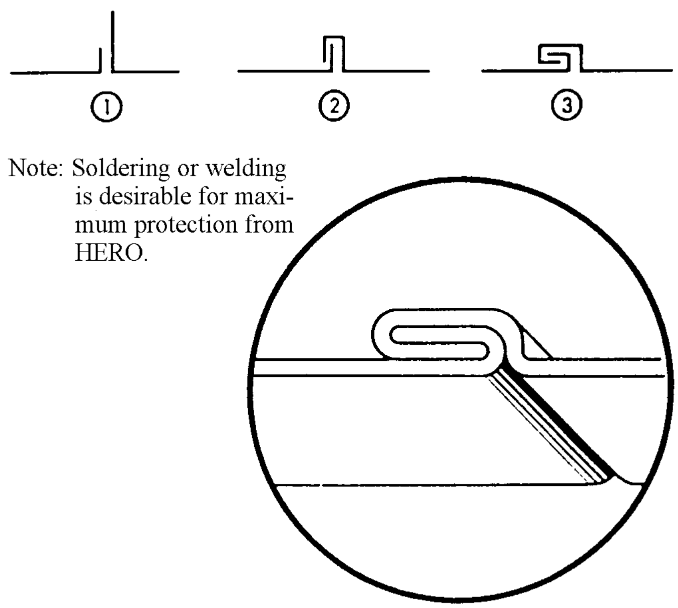


Figure 5-21. Formation of Permanent Crimp Seam

c. When protective coatings are necessary, they should be so designed that they can be easily removed from mating surfaces prior to bonding. Since the mating of bare metal to bare metal is essential for a satisfactory bond, a conflict may arise between the bonding and finish specifications. From the viewpoint of shielding effectiveness, it is preferable to remove the finish where compromising of the bonding effectiveness would occur.

d. Certain protective metal platings such as cadmium, tin, or silver need not in general be removed. Similarly, low-impedance corrosion-resistant finishes suitable for aluminum alloys, such as alodine, iridite, oakite, turco and bonderrite, may be retained. Most other coatings, such as anodizing, are nonconductive and would destroy the concept of a bond offering a low impedance radio frequency path.

e. Mating surfaces should be bonded immediately after protective coatings are removed to avoid oxidation. Refinishing after bonding is acceptable from the standpoint of shielding effectiveness.

f. When two dissimilar metals must be bonded, metals that are close to one another in the electro-chemical series should be selected.

g. Soldering may be used to fill the resulting seam, but should not be employed to provide bond strength.

h. The most desirable bond is achieved through a continuous butt or lap weld. Spot-welding is less desirable because of the tendency for buckling, and the possibility of corrosion occurring between welds. Riveting or pinning is even less desirable because of the greater susceptibility of bond degradation with wear.

i. An overlap seam, accompanied by soldering or spot welding, provides a relatively effective bond. Other types of crimped seams may be employed, so long as the crimping pressure is maintained.

There are often occasions when good temporary bonds must be obtained. Bolts, screws, or various types of clamp and slide fasteners have been used for this purpose.

The same general requirements of clean and intimate contact of mating surfaces, and minimized electrolytic (cathodic) effects apply to temporary bonds as well. Positive locking mechanisms that ensure consistent contact pressure over an extended period of time should be used. Lockwashers should be employed that can "bite into" metal surfaces and fasteners and maintain a low bonding resistance. Bolts, nuts, screws, and washers that must be manufactured with material different from the surfaces to be bonded should be higher in the electromotive series than the surfaces themselves, so that any material migration erodes replaceable components.

A critical factor in temporary bonds (and in spot-welding permanent bonds as well) is the linear spacing of the fasteners or spot-welds. Figure 5-22 provides an indication of the sensitivity of this parameter for a 2-inch aluminum lap joint. Data are taken at 200 MHz. The shielding effectiveness shown at 1-inch spacing is about 12 dB poorer than the identical configuration incorporating a 2-inch wide monel mesh gasket; the effectiveness at 10-inch spacing is about 30 dB poorer than the same gasket. Use of conductive gaskets for this and other applications will be discussed shortly.

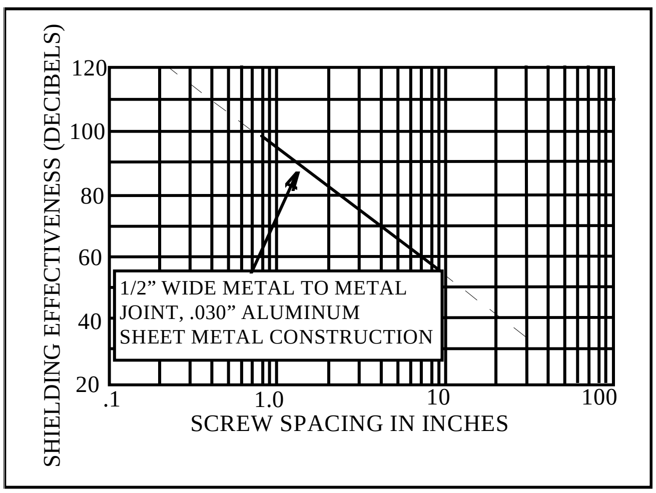


Figure 5-22. Influence of Screw Spacing

5-6.2. SEAMS WITH GASKETS. Considerable shielding improvement over direct metal-to-metal mating of shields used as temporary bonds can be obtained using flexible, resilient electrically conductive gaskets placed between shielding surfaces to be joined. Clean conductive mating surfaces and a good pressure contact are necessary. Figures 5-23 through 5-26 are good examples of this type of bonding.

The major material requirements for RF gaskets include compatibility with the mating surfaces, corrosion resistance, appropriate electrical properties (refer to table 5-1), resilience (particularly when repeated compression and decompression of the gasket is expected), mechanical wear, and ability to form into the desired shape. On this basis, monel and silver-plated brass are generally the preferred materials, with aluminum surfaces. Beryllium-copper contact fingers are usually employed, with a variety of platings available, if desired. Mumetal and Permalloy have been used when magnetic shielding effectiveness is a concern.

For applications requiring moisture/pressure sealing as well as RF shielding, combination rubber-metal seals are available. These include metal mesh bonded to neoprene or silicone, aluminum screen impregnated with neoprene, oriented wires in silicone, conductive adhesives and sealants, and conductive rubber. The advantages and limitations of these, as well as non-sealing RF gaskets, are summarized in table 5-10.

The problem in the marine environment of salt water/salt spray is that the fillers of conductive composite gaskets may corrode. For applications requiring corrosion resistance, special composite gaskets which avoid this may be used. One possible solution is to protect the conductive fillers in the matrix, with an organic coating. If a metal mesh or knit is used, Raychem GELTEK (a silicon gel) may be used to impregnate the mesh or knit. This gel protects the conductive material from the corrosive elements.

Similar to matching dissimilar metals to be bonded (see paragraph 5-6.1), the fillers of composite gaskets should be galvanically similar to the metal surfaces they join (i.e., close to each other in the electrochemical series).

Knitted wire mesh gaskets can be manufactured in a variety of shapes and sizes. Gasket thickness is dependent on the unevenness of the joint to be sealed, the compressibility of the gasket, and the force available. Gasket shape depends on the particular application involved, as well as the space available, the manner in which the gasket is held in place, and the same parameters that influence gasket thickness.

5-6.3. TEMPORARY APERTURES AS DISCONTINUITIES. Temporary apertures of a weapon are those apertures, such as access panels, that must be open during adjustment or installation of circuits or components. They should be designed so that when they are closed, a low RF impedance electrical bond is maintained between the door or panel and the weapon housing. The best way of achieving this is to use metallic gaskets or finger stock between the mating surfaces. When metallic finger stock is used, 5 to 10 grams of pressure per finger should be applied to the mating surfaces.

If hinges are used on panels, it is recommended that gasketing such as conductive weather stripping be used on the hinged side of the panel. An alternative method of shielding at the hinge side of a panel is to use metal finger stock. The shielding material must be electrically and mechanically bonded to the frame at close intervals to ensure proper shielding.

Figure 5-27 illustrates acceptable methods of applying shielding materials around the sides of hinged access panels. Appropriate mechanical locking devices must be used on access panels to maintain a minimum of 20 psi pressure between the panel and the gasket or fingers.

The best arrangement of spring contact fingers around removable panels or doors calls for the installation of two sets of fingers at right angles to each other. One set is a wiping set, the other is in compression, and the combination makes good electrical contact when the door is closed. The pressure exerted by these springs is highly important and it should be carefully maintained.

Access panels or doors cannot perform a shielding function when opened or removed. If it is necessary for apertures to be opened in EM fields, the interior circuits, components, and cables should be designed to preclude HERO.

Silver-filled silicon rubber gaskets can be obtained in sheet, die-cut, molded, or extruded form. The most popular, and most economical of these types, is the extrusion. Figure 5-28 shows typical extruded shapes and indicates recommended deflection limits for various shapes and sizes. Comments made above concerning thickness, shape, and mounting methods for wire mesh gaskets also apply to conductive rubber gaskets.

Table 5-10. Characteristics of Conductive Gasketing Materials

MATERIAL	CHIEF ADVANTAGES	CHIEF LIMITATIONS
Compressed knitted wire	Most resilient all-metal gasket (low flange pressure required). Most points of contact. Available in variety of thicknesses and resiliencies, and in combination with neoprene and silicone.	Not available in sheet (certain intricate shapes difficult to make). Must be 0.040 in. or thicker. Subject to compression set.
Brass or beryllium copper with punctured nail holes	Best break-thru of corrosion protection films.	Not truly resilient or generally reusable.
Oriented wires in rubber silicone	Combines fluid and RF seal. Can be effective against corrosion films if ends of wires are sharp.	Might require wider or thicker size gasket for same effectiveness. Effectiveness reduces with mechanical use.
Aluminum screen impregnated with neoprene	Combines fluid and conductive seal. Thinnest gasket. Can be cut to intricate shapes.	Very low resiliency (high flange pressure required).
Soft metals	Cheapest in small sizes.	Cold flows, low resiliency.
Metal over rubber	Takes advantage of the resiliency of rubber.	Foil cracks or shifts position. Generally low insertion loss yielding poor RF properties.

Table 5-10. Characteristics of Conductive Gasketing Materials (Continued)

MATERIAL	CHIEF ADVANTAGES	CHIEF LIMITATIONS
Conductive rubber (carbon filled)	Combines fluid and conductive seal.	Provides moderate insertion loss.
Conductive rubber (silver filled)	Combines fluid and RF seal. Excellent resilience with low compression set. Reusable. Available in any shape or cross-section.	Not as effective as metal in magnetic fields. May require salt spray environmental protection.
Contact fingers	Best suited for sliding contact.	Easily damaged. Few points of contact.

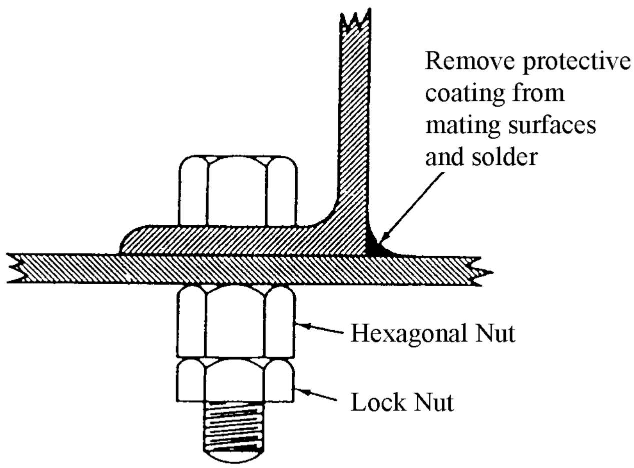


Figure 5-23. Acceptable Bonding Technique Using Bolts

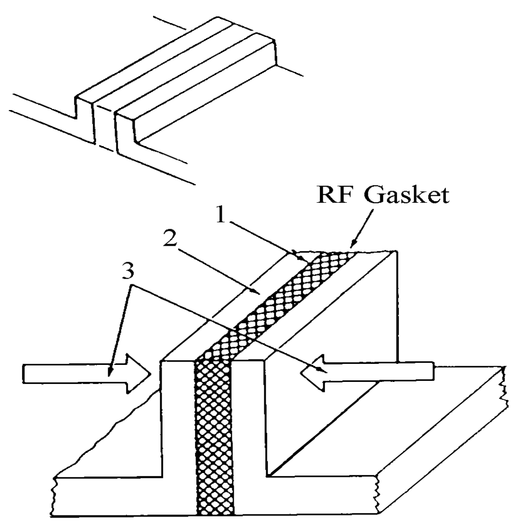


Figure 5-24. Acceptable Method of Making Permanent Seam Using RF Gasket

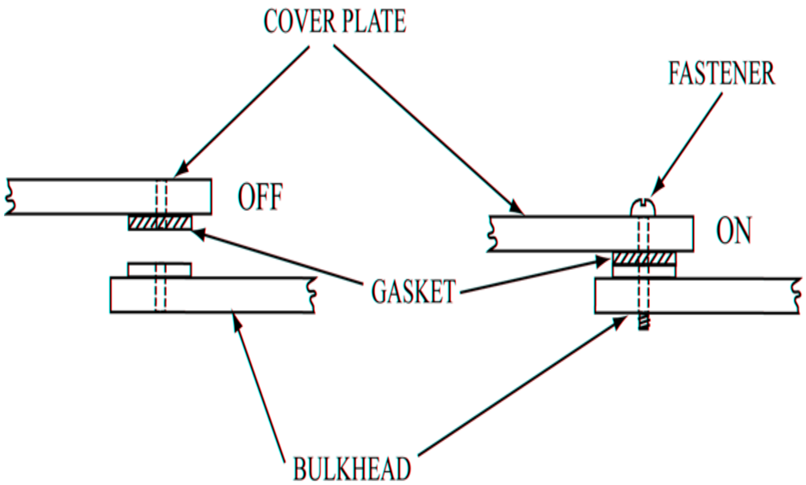


Figure 5-25. Cover Plates with Gaskets

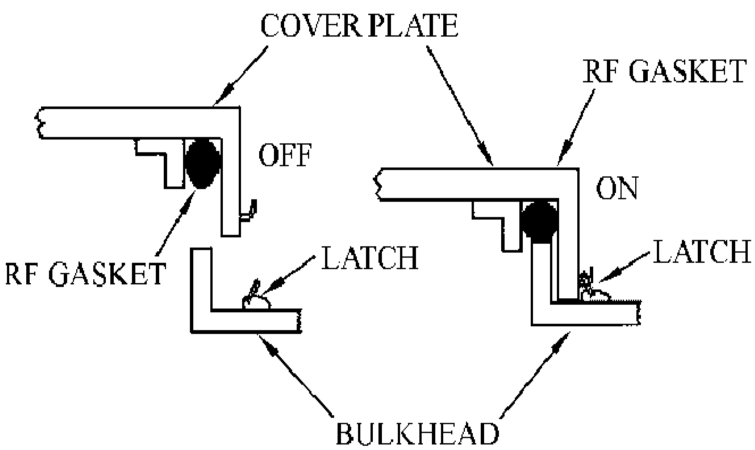


Figure 5-26. Covers with Gaskets

Shielding effectiveness of silver-filled (or silver-plated copper filled) silicones is especially high between 15 kHz and 10 GHz. Plane wave attenuation often improves with higher closure force, especially for die-cut gaskets. Best results are achieved with molded or extruded cross-sections held in grooves.

Gaskets may be held in place by sidewall friction, by soldering, by adhesive, or by positioning in a slot or on a shoulder. Soldering must be controlled carefully to prevent its soaking into the gasket and destroying gasket resiliency. Adhesives (particularly non-conductive adhesives) should not be applied to gasket surfaces that mate for RF shielding purposes; auxiliary tabs should be used.

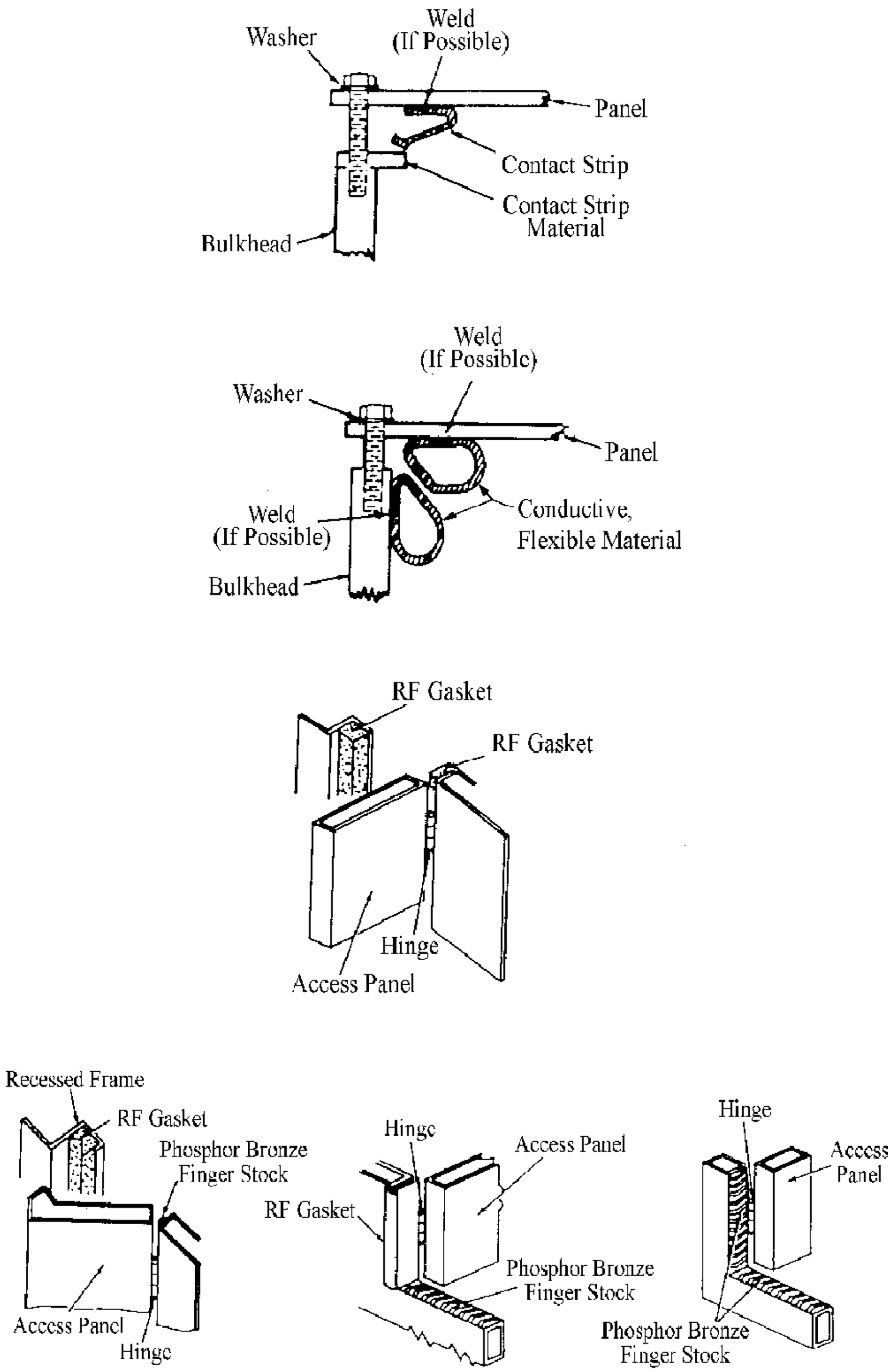
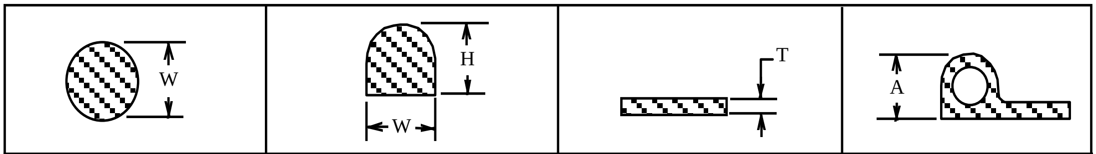


Figure 5-27. Acceptable Methods for Temporary Aperture Design



Deflection	W Dia.	Deflection	H	Deflection	T	Deflection	A
.007 – .018	.070	.006 – .012	.068	.001 – .002	.020	.025 – .080	.200
.010 – .026	.103	.008 – .016	.089	.001 – .003	.032	.030 – .125	.250
.013 – .031	.125	.012 – .024	.131	.003 – .006	.162	.075 – .250	.360
.014 – .035	.139	.014 – .029	.156	.003 – .009	.093		
		.016 – .032	.175				

from “EMI/RFI Gasket Design Manual,” 1975, Chromerics, Inc.

Figure 5-28. Gasket Deflection Limits (In Inches)

Typical gasket pressures for obtaining effective seals range from 5-100 psi. A usual pressure is 20 psi.

Five of the most-often encountered RF gasket design problems are illustrated in figure 5-29, along with the means of overcoming the problems.

A practice presently in use is the dating of conductive and pressure gaskets, as well as other items that may suffer wear with continued use. The date is used during maintenance as an indicator of whether the gasket or other tagged item should be replaced.

5-6.4. USE OF WAVEGUIDE ATTENUATORS. The effects of seams, cracks, openings, holes, and other breaks in a shield can often be studied by considering the opening as a waveguide. Alternatively, waveguide attenuators can be designed around control shafts, light receptacles, and other openings to reduce interference leakage.

If the opening can be treated as a rectangular aperture, its cutoff wavelength, λ_c, is

$$\lambda_c = 2W \tag{5-19}$$

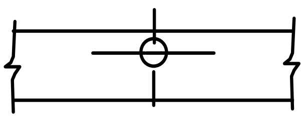
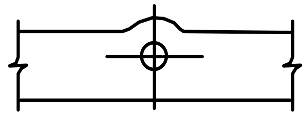
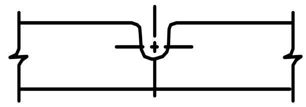
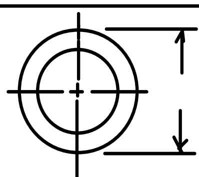
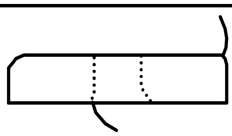
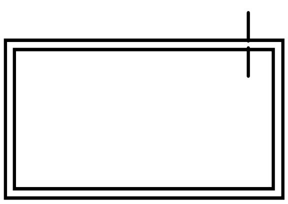
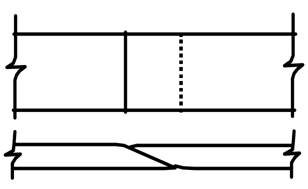
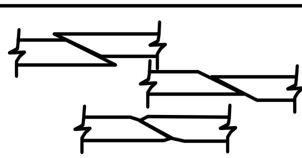
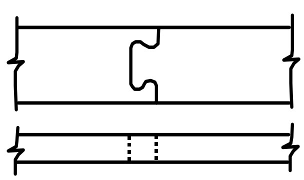
detail	why faulty	suggested remedy
 Bolt holes close to edge	Causes breakage in stripping and assembling	Projection or “ear”  Notch instead of hole 
 Metalworking tolerances applied to gaskets thickness, diameters, length, width, etc.	Results in perfectly usable parts being rejected at incoming inspections, Requires time and correspondence to reach agreement on practical limits. Increases cost of parts and tooling. Delays deliveries	Most gasket materials are compressible. Many are affected by humidity changes. Try standard or commercial tolerances before concluding that special accuracy is required
 Transference of fillets, radii, etc. from mating metal parts to gasket	Unless part is molded, such features mean extra operations and higher cost	Most gasket stocks will conform to mating parts without pre-shaping, Be sure radii, chambers, etc. are fuctional, not merely copied from metal
 The walls delicate cross section in relation to over-all size	High scrap loss: stretching or distortion in shipment or use. Restricts choice to high tensile strength materials.	Have the gasket in mind during early design stages
 Large gaskets made in sections with beveled joints	 Extra operations to skive. Extra operations to glue. Difficult to obtain smooth, even joints without steps or transverse grooves	 Die-cut dovetail joint

Figure 5-29. Common Errors in Gasket Design

where

w = the long dimension of the opening.

Slot attenuation, in dB, can be computed from the relationship

$$A = 54.5d/\lambda_c\sqrt{\frac{1-\lambda_c}{\lambda}} \qquad (5\text{-}20)$$

where

A = the attenuation per unit depth

d = slot depth

λ = interference wavelength

Similar relationships can be established for other opening configurations.

A useful rule to follow for circular holes is that, for 100 dB of attenuation, the length of the waveguide must be at least three times the diameter of the hole. The maximum permissible diameter (d_{max}) can be obtained by dividing the wavelength for the highest frequency under consideration (λ_{min}) by 3.4, where both diameter and wavelength are expressed in the same units.

$$d_{max} = \lambda_{min}/3.4 \qquad (5\text{-}21)$$

In many cases, shielding screens introduce excessive air resistance and sometimes greater shielding effectiveness may be needed than they can provide. In such cases, openings may be covered with specially designed ventilation panels (such as honeycomb) with openings that operate on the waveguide-below-cutoff principle. Honeycomb-type ventilation panels in place of screening:

a. Allow higher attenuation than can be obtained with mesh screening over a specified frequency range.

b. Allow more air to flow without pressure drop for the same diameter opening.

c. Cannot be damaged as easily as the mesh screen, and are therefore more reliable.

d. Are less subject to deterioration by oxidation and exposure.

All non-solid shielding materials, such as perforated metal, fine mesh copper screening, and metal honeycomb, present an impedance to air flow. Metal honeycomb is the best of these materials because it enables very high electric field attenuations to be obtained through the microwave band with negligible drops in air pressure. However, honeycomb has the disadvantages of occupying greater volume and costing more than screening or perforated metal. Also, it is often difficult to apply honeycomb paneling because flush mounting is required. Thus, screening and perforated sheet stock sometimes find application for purely physical design reasons, although honeycomb panels can achieve attenuations to 136 dB above 10 MHz.

5-6.5. PANEL OPENINGS. Panel openings that must accommodate control shafts may be shielded in one of several ways. A waveguide attenuator may be used around the panel opening as discussed under paragraph 5-6.4, so long as the shaft within the guide is non-conducting. Alternatively, the portion of the control that is behind the panel may be shielded to separate the control from the remainder of the equipment, and the control leads filtered.

A military requirement exists regarding RFI-shielded rotary-shaft seals (MIL-B-5423/16C). The requirement specifies a brass, nickel-plated nut with a knitted monel wire mesh insert to ground the control shaft.

Fuseholders, phone jacks, panel connectors not in use, and other receptacles can be fitted with a metallic cap that provides an electrically continuous cover and maintains case integrity.

Pushbutton switches are recommended to be shielded using boots having a mesh gasket insert (MIL-B-5423/7A). This approach appears to be more effective than the alternative of a conductive rubber boot.

5-6.6. REQUIRED VISUAL OPENINGS. Often, it is necessary to provide RF shielding over pilot-light bulbs, digital display faces, meter faces, strip chart recorder outputs, oscilloscope faces, or similar devices that must be observed by the equipment user. The alternatives available include the following:

a. Use of a waveguide attenuator;

b. Use of screening material;

c. Providing a shield behind the assembly of concern, and filtering all leads to the assembly; and

d. Use of conducting glass.

A waveguide attenuator is a practical approach for RF shielding of lamps. This technique has the advantage of not introducing light transmission loss. However, it is not particularly suitable for most meter openings or larger apertures, because of the space requirements involved.

Use of screens over meter faces and other large apertures has often been employed for shielding purpose. A typical screen introduces a minimum of 15-20 percent optical loss, and can create difficulties in reading meters. If the device being shielded has a scale (such as an oscilloscope reticule), bothersome zoning patterns can result. However, these potential deficiencies are counterbalanced by good shielding efficiencies at a fairly low cost.

Figure 5-30 illustrates one method of mounting such screens, when they are not incorporated directly into the device to be shielded. The screen may be embedded in the center of a single pane of acrylic, or incorporated into a glass sandwich. It should be tinned or otherwise bonded around its periphery to achieve good mating to the metallic plate. A variety of screen materials are available. Slight variations in the dimensions of the mesh openings are intentionally made to reduce the meter-reading and zoning problems.

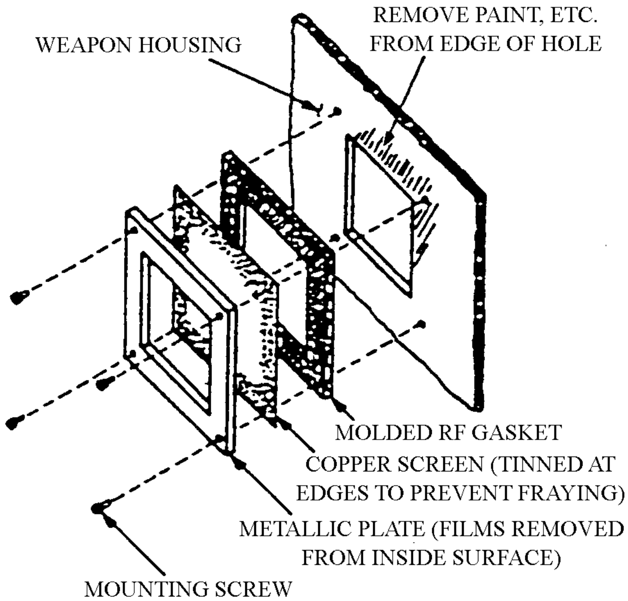


Figure 5-30. Method of Mounting Wire Screen over a Large Aperture

Two approaches can be employed when shielding behind an assembly and then filtering all leads passing through the shield. These approaches are shown in figure 5-31A and 5-31B, in the case of a meter. One method, when the meter involved is essentially an off-the-shelf item, is to build a supplementary enclosure and to pass the meter leads through feed-through capacitors or other appropriate filters to eliminate interference that may have been picked up through the meter face. The other method is to procure a meter whose back can be used in place of a supplementary shield, and one which incorporates the necessary lead filtering. In either case, it is assumed that external fields will not cause adverse effects to the operation of the meter itself.

Glass coated with conducting material such as silver can provide shielding across viewing surfaces with some loss in light transmission. Conductive glass is commercially available from a number of glass manufacturers.

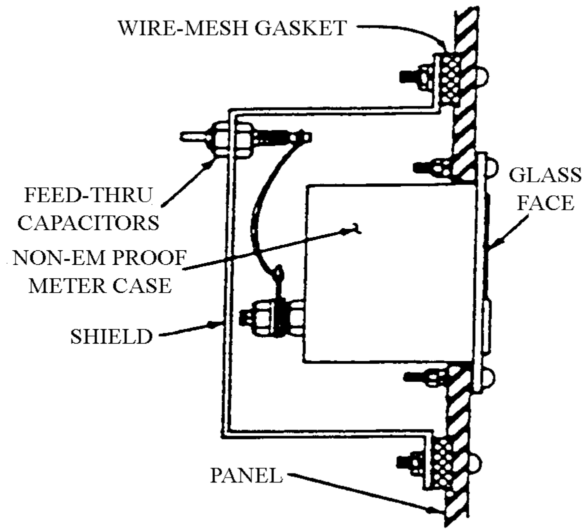


Figure 5-31A. Meter Shielding Techniques

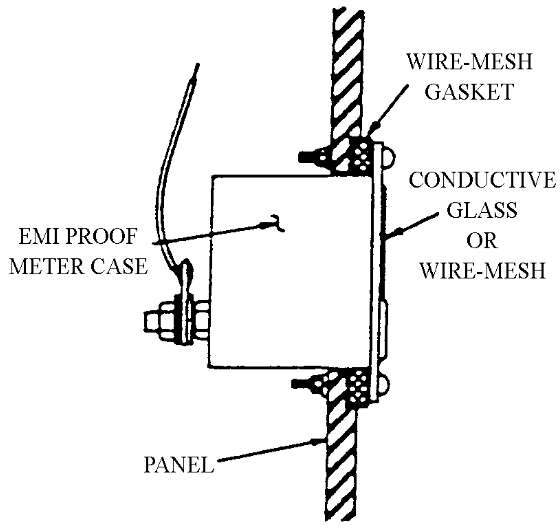


Figure 5-31B. Shielded Meter Techniques

The light transmission characteristics of this type of glass as a function of surface resistance are presented in figure 5-32. For effective shielding, good contact to the conducting surface of the glass must be maintained around its periphery.

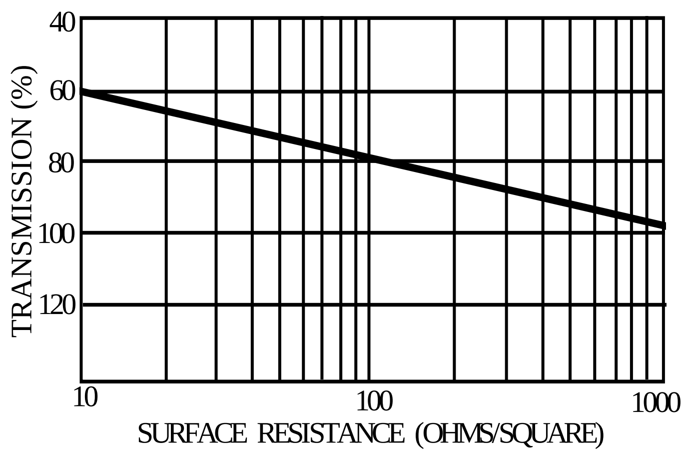


Figure 5-32. Light Transmission Vs Surface Resistance for Transparent Conductive Glass

5-7. SHIELDING TEST METHODS.

5-7.1. GENERAL. The variety of test methods available for evaluating shielding effectiveness is due, at least in part, to the many different factors that can affect material shielding capabilities. These factors include the configuration of the shield (is it a sheet of material, or is it a box?), the frequency range of concern, whether or not the impinging wave is planar, the wave impedance, and others.

This section identifies the most frequently employed and generally applicable shielding effectiveness tests. While they are not all appropriate to evaluating a particular design, they provide a set of established procedures from which designers can select applicable techniques. These tests include:

a. Low-Impedance Magnetic Field Testing Using Small Loops,

b. Low-Impedance Magnetic Field Testing Using a Helmholtz Coil,

c. High-Impedance Electric Field Testing Using Rod Antennas,

d. High-Impedance Electric Field Testing Using a Parallel Line Radiator,

e. Plane Wave Testing Using Antenna,

f. Plane Wave Testing Using a Parallel Plate Transmission Line,

g. MIL-STD-1377 Testing, and

h. Mode Stirred Chamber Testing.

The MIL-STD-1377 tests represent recommended procedures for evaluating the shielding (and filtering) effectiveness of Navy weapons systems. Although this standard has not been updated, it contains a unique approach to shielding measurements, and its cable effectiveness evaluation methods are good illustrations of how cable and connector performance tests can be performed.

A number of the above tests are very similar to tests designed to measure equipment and system EMC in accordance with MIL-STD-462. They also are similar to tests performed to evaluate EM effectiveness of shielded enclosures used for testing purposes in accordance with MIL-STD-285. The design engineer who is concerned with the measurement of shielding properties should be familiar with both of these standards.

The MIL-STD-1377 tests represent recommended procedures for evaluating the shielding (and filtering) effectiveness of Navy ordnance systems. The specification contains a unique approach to shielding measurements, and its cable effectiveness evaluation methods are good illustrations of how cable and connector performance tests can be performed.

5-7.2. MIL-STD-1377 TESTING. This standard is employed by weapon system developers to evaluate shielding and filtering effectiveness. It defines applicable shielding effectiveness measurement techniques covering the frequency ranges of 100 kHz to 30 MHz, and 1000 MHz to 10 GHz; and filtering effectiveness measurement techniques covering the frequency range of 100 kHz to 10 GHz.

At frequencies below 30 MHz, shielding in MIL-STD-1377 is evaluated on the basis of <u>Surface Transfer Impedance</u> (STI). STI is the ratio of the longitudinal voltage drop on the outer surface of the shield to the current creating that voltage drop. If a cable or cable/connector is under test, it is terminated in a short, and driven by a signal source. The STI is measured with an RF voltmeter. Shield enclosure discontinuities are measured in a similar way. An STI of about 10 milliohms has been correlated to a level of 15 percent of maximum no fire current for typical EID's.

This standard, while outdated, does provide effective test methods over the test frequencies specified in the standard. New requirements covering the frequency range 30 MHz to 1000 MHz require HERO tests over this range. The primary method used to cover this frequency range uses the mode-stirred chamber technique which provides worst-case response measurements of the ordnance systems when subjected to exposure in the chamber. The test methods of MIL-STD-1377 used in the 1 GHz to 10 GHz range can be applied up to 40 GHz for tests requirements which include these extended frequency bands.

5-8. SUMMARY OF GOOD SHIELDING PRACTICES.

The following represent what might be considered the more salient points on ordnance system shielding design considerations covered or implied in the previous discussion:

a. Good conductors such as copper, aluminum, and magnesium should be used for high-frequency electric-field shields to obtain the highest reflection loss.

b. Magnetic materials such as iron and mumetal should be used for low-frequency, magnetic-field shields to obtain the highest penetration loss.

c. Any shielding material strong enough to support itself will usually be thick enough for shielding electric fields at any frequency.

d. In the case of thin-film shields, the effectiveness of the shield is fairly constant for material thicknesses below $\lambda/4$, and increases markedly above that thickness.

e. Multiple shields (for both enclosures and cables) can provide both higher shield effectiveness and extended shielding frequency range. Cost considerations will probably be the deciding factor between use of multiple shields and using other means of achieving HERO protection, although factors such as reduced cable flexibility with double braids may also come into play.

f. All openings or discontinuities should be treated in the design process, to assure minimum reduction in shield effectiveness. Particular attention should be paid to selection of materials that are not only suitable from the shielding standpoint, but from the electro-chemical corrosion viewpoint as well.

g. When other aspects of system design will permit, continuous butt or lap weld seams are most desirable. The important consideration is to get intimate contact between mating surfaces over as much as the seam surfaces as possible.

h. Surfaces to be mated must be clean and free from nonconducting finishes unless the bonding process positively and effectively cuts through the finish. When EMC and finishing specifications conflict, it is important that the finishing requirement be modified.

i. The critical factors in weapon system cable shielding are shield coverage under operational cable flexing conditions, and cable shield termination at the connector. A minimum of 94 percent shield coverage is recommended for these applications. Shields must be peripherally bonded to connector back shells to maintain shielding effectiveness at mating surfaces.

j. Conductive gaskets and spring fingers, waveguide attenuators, screens and louvers, and conducting glass are the major devices and mechanisms available for maintaining enclosure shielding effectiveness. Many factors, in addition to shielding capabilities per se, ranging from space availability to cost, and from air circulation requirements to visibility factors, will affect particular methods employed in particular situations.

k. Shielding represents only one method of reducing ordnance system susceptibility, and should not be considered without also considering tradeoffs of filtering, grounding, and bonding techniques which complement the use of shields.

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CHAPTER 6

BONDING

6-1. GENERAL.

Bonding is the establishment of a low-impedance path between two metal surfaces. This path may be between two points on a system ground plane, or between ground reference and a component, a circuit, or a structural element. The purpose of the bond is to make the structure homogeneous with respect to the flow of radio-frequency (RF) currents, thus avoiding the development of electric potentials between metallic parts which can produce interference.

Generally there are two types of bonding: direct bonding, where there is metal-to-metal contact between the members to be bonded; and indirect bonding through the use of conductive jumpers.

Permanent joints of metallic parts made by welding, brazing, sweating, swaging, and soldering are the best direct bonds. Semi-permanent joints of machined metallic surfaces rigidly held together provide excellent direct bonds, as long as the contact areas are clean and all non-conductive coatings are removed prior to assembly. Joints that are press-fitted or jointed by self-tapping or sheet metal screws cannot be relied upon to provide a low-impedance bond at high frequencies. Riveted joints on $^{3}/_{4}$" centers are acceptable if the rivet holes are bare. Direct bond must always be made through continuous contact to base or conductively finished metals.

An indirect bond (or bonding jumper or strap) is an intermediate electrical conductor used to connect two isolated items. Because jumpers often have significant impedance at frequencies above 10 MHz, direct bonds are preferable. However, there are obviously circumstances when direct bonds cannot be used (such as when equipment must be removable, or when mechanical shock-mounting is necessary); and these situations warrant indirect bonds.

Good bonding between equipment and a ground reference plane is essential to minimizing interference because:

a. Good bonding enables the design objectives of other methods of electromagnetic interference (EMI) suppression, such as shields and filters, to be more nearly achieved.

b. Good bonding minimizes the build-up of RF voltage differences and ground current loops.

c. An adequate bond deters the build-up of static charges in equipment operation.

In addition to reducing interference, good bonding minimizes damage which might be caused by lightning strikes, and protects personnel from the shock hazard that could result if primary power were inadvertently shorted to an enclosure.

An example of the effects of a poor bond is shown in figure 6-1. If the impedance of the bond is significant, the interference signal, which should be bypassed by the filter capacitors (Path 1), will be coupled into the susceptible equipment (Path 2).

L_B= INDUCTANCE OF A POOR BOND

R_B= CONTACT RESISTANCE OF A POOR BOND

NOTE - WHEN $(I/J\omega C + R_L) < (R_B + J\omega L_B)$ INTERFERENCE CURRENTS WILL FOLLOW PATH (2) TO SUSCEPTIBLE EQUIPMENT.

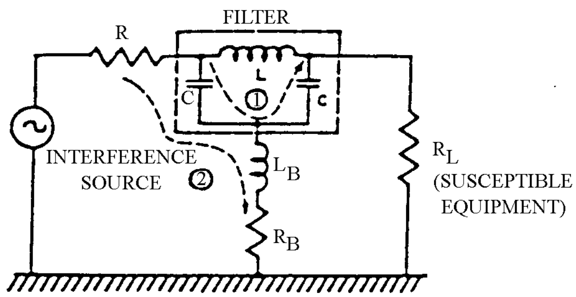


Figure 6-1. Circuit Representing Poor Bonding between a Filter and Ground

When implementing bonding techniques, it must always be remembered that bonding straps do not provide a low-impedance current path at RF. The impedance important in this discussion is the impedance at RF. There is little correlation between the dc resistance of a bond and its RF impedance. Even the measured RF impedance of bonds, such as jumpers, straps, rivets, etc., is not a reliable indication of the bonding effectiveness in the actual installation. It should also be remembered that conductive epoxies and pastes are not always sufficient RF bonds. Even when proven effective in certain instances, they have been known to degrade shielding effectiveness under conditions of strain, pressure, and the passage of time.

6-2. BONDING DESIGN GUIDELINES.

The effectiveness of a bond depends on its application, frequency range, magnitude of current, and environmental conditions such as vibration, temperature, humidity, fungus, and salt content of the ambient environment. Figures 6-2 through 6-6 provide examples of direct and indirect bonds, and illustrate the variety of techniques that have been employed to obtain low-impedance connections. Many other examples are available, particularly in MIL-STD-1310.

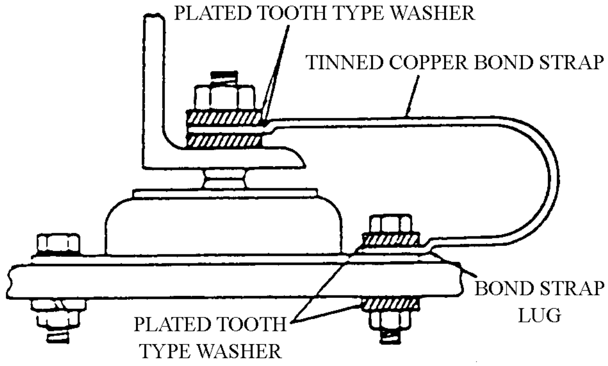


Figure 6-2. Typical Shock Mount Bond

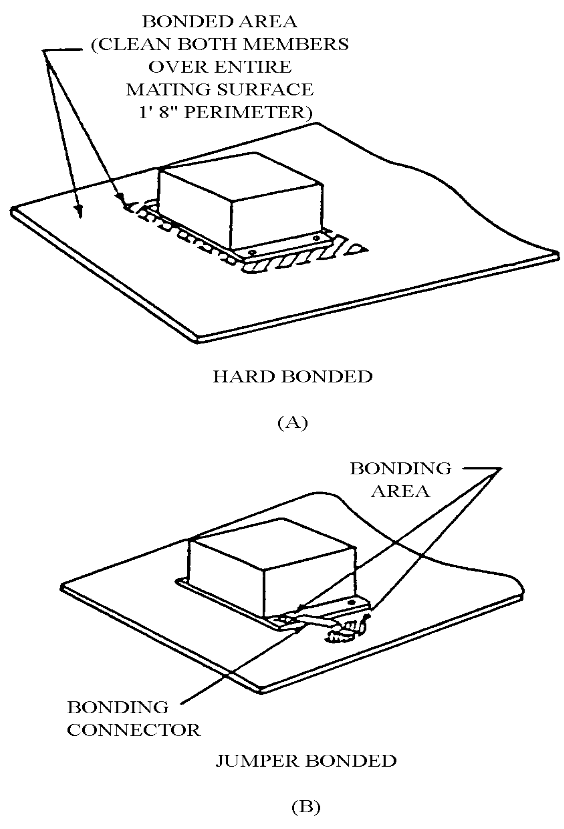


Figure 6-3. Base Components

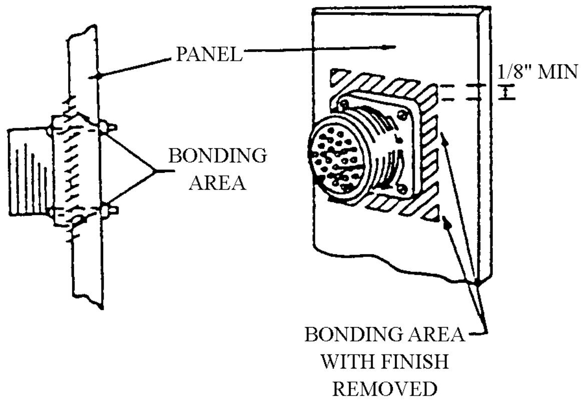


Figure 6-4. Bonding of a Connector

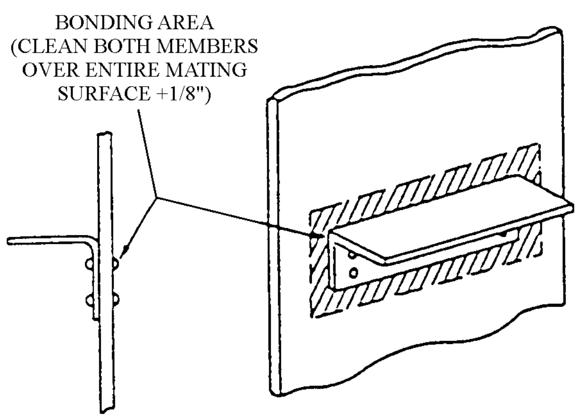


Figure 6-5. Bolted Members

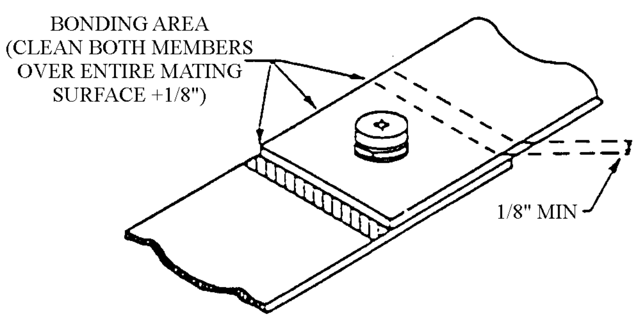


Figure 6-6. Bracket Installation (Rivet or Weld)

Some general guidelines for obtaining good bonds are provided in the following paragraphs.

Good bonding depends on intimate contact between metal surfaces. Surfaces must be smooth, clean, and free of non-conductive finishes. The fastening method must exert sufficient pressure to hold the surfaces in contact in the presence of deforming stresses, shock, and vibrations associated with the equipment and its environment.

The best bonds are always made by joining similar metals. If this is not possible, special attention must be paid to the possibility of bond corrosion, the choice of the materials to be bonded, the selection of supplementary components (such as washers) that will assure any corrosion will affect replaceable elements only. Solder should not be used to provide mechanical strength to a bond. Protection of the bond from moisture and other corrosion effects must be provided where necessary.

Bonding jumpers are only a substitute for direct bonds. The jumpers should be kept as short as possible, have a low resistance and low L/C, and not be lower in the electro-chemical series than the bonded members. A good rule to use is that the jumper should have a length-to-width ratio of less than 5.

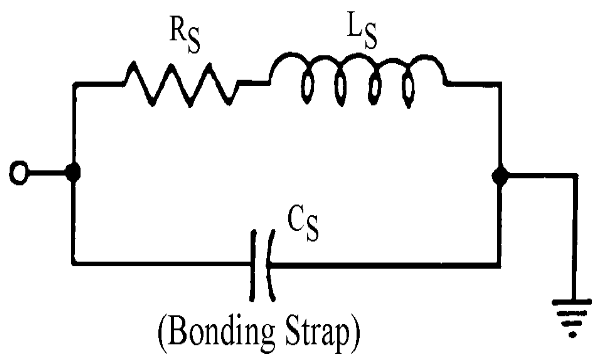
Jumpers should be bonded directly to the basic structure, rather than through an adjacent part. They should not be connected with self-tapping screws, or by any other means where screw threads are the primary means of bonding.

It is always important in the broadest types of bonding application that the bonding jumper or direct bond is sufficient to carry the currents that may flow through it. Use single-point bonds at low frequencies (circuit dimensions $\lambda / 20$) and multi-point bonds at high frequencies.

6-3. BONDING EFFECTIVENESS CHARACTERISTICS.

6-3.1. BOND JUMPER EQUIVALENT CIRCUIT. The use of bonding jumpers in indirect bonding is equivalent to the problem of maintaining low-impedance paths. At low frequencies, bonding jumpers do not present any special problems except resistance, and any reasonable length jumper can be used. At higher frequencies, however, the RF impedance of the bond becomes a critical design consideration.

The equivalent circuit of a bond strap at frequencies where the length of the strap is short compared to a wavelength is a simple parallel-tuned circuit, as illustrated in figure 6-7A. As such, the bonding jumper has the usual electrical parameters of resistance, inductance, and capacitance. Of these parameters, its resistance is an inherent property of jumper resistivity, depending on the material selected; its capacitance is dependent upon the physical configuration and separation from the bonded units; and its inductance is dependent upon the physical dimensions of the bonding jumper.



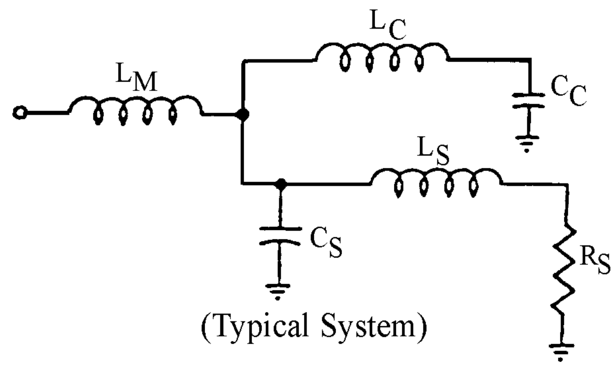


Figure 6-7A. Bonding Equivalent Circuit at Low Frequencies

Figure 6-7B. Bonding Equivalent Circuit at High Frequencies

The dc resistance of a bonding strap per unit length of the bond can be obtained from the relationship

$$R_{dc}\ (ohm/cm) = p/A \tag{6-1}$$

where

ρ = the specific resistivity of the material, and

A = the cross-sectional area of the bond.

If ρ is in units of ohm-centimeters (for copper, $\rho = 1.724 \times 10^{-6}\ ohm - cm$) and A is in square-centimeters, then R_{dc} is in ohms-per-centimeter of bond length.

At frequencies where skin effect becomes significant, the ac resistance of the bond can differ significantly from its dc value. Skin effect is defined as the depth at which the current density is 1/e (about 37%) of its surface current density. This depth, in centimeters, is defined as:

$$\delta = 5033\sqrt{\rho/\mu f}, \tag{6-2}$$

where

f = the frequency, in Hertz

μ = material permeability, relative to copper

The assumption that all of the current in a conductor is within the skin depth defined by equation (6-2) results, in the case of circular conductors, in the following equation for ac resistance:

$$R_{ac}(ohms/cm) = \frac{\rho}{2\pi r\delta}, \quad (6\text{-}3)$$

where

r = the conductor radius, in centimeters.

Figure 6-8 indicates the ratio of ac-to-dc resistance for several sizes of copper wire over the frequency range of 1-1000 MHz.

For a straight bonding strap of non-magnetic metal, bonding inductance, L, is defined as:

$$L = 0.00508\,a[2.303LOG_{10}(2a/b + c) + 0.5 + 0.2205\,(b + c/a)] \quad (6\text{-}4)$$

If the bond is made with a straight piece of wire of circular cross section, L is given by:

$$L = 0.00508\,a[2.303LOG_{10}(4a/d) - 0.75] \quad (6\text{-}5)$$

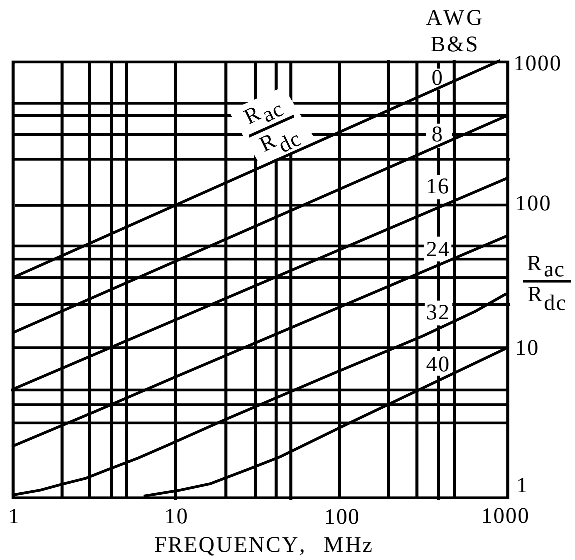


Figure 6-8. Ratio of AC to DC Resistance for Various Wire Sizes

in the last two equations,

a = the length of the strap in inches

b = the width of the strap in inches

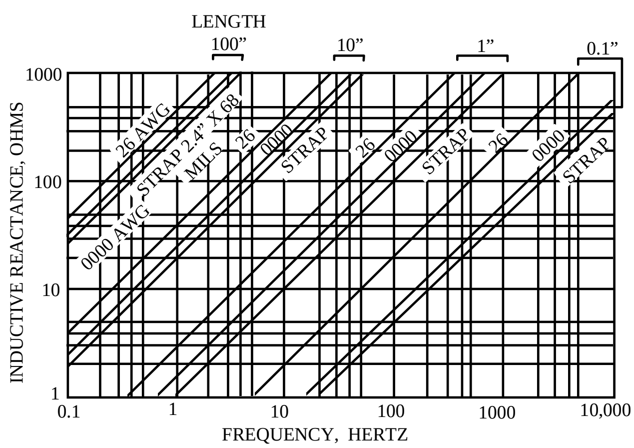
c = the thickness of the strap in inches, and

d = the wire diameter in inches

Figure 6-9 is a plot of the calculated inductive reactance of three specific bonds as a function of frequency. Two of the bonds are copper wire, while the third is a copper strap.

At frequencies where the length of the bond approaches or exceeds the wavelength of concern, the bond can act as a transmission line. The result is the development of standing-waves on the bond, and a deterioration in bond effectiveness.

Typical bond straps may be manufactured as flat, solid straps; braided straps; or solid wire. The material used is generally copper or aluminum, but in the case of flat straps, it can also be phosphor bronze.



CALULATED INDUCTIVE REACTANCE, IN UH, FROM,

$L = 0.00508\, a\, [2.303\, \mathrm{LOG}_{10}\, (4a/d) - 0.75]$, OR

$L = 0.00508\, a\, [2.303\, \mathrm{LOG}_{10}\, (2a/b+c) + 0.5 + 0.2205(b+c/a)]$

WHERE a = Length, (inches)
b = Width, (inches)
c = Thickness, (inches)
d = Diameter, (inches)

Figure 6-9. Inductive Reactance of Wire and Strap Bond Jumpers

6-3.2. EQUIPMENT EFFECTS ON INDIRECT BONDS. The representation of the equivalent circuit of a bond as a parallel-tuned circuit ignores the effects of the equipment enclosure or other item being bonded, and the latter can significantly affect bond strap performance. Figure 6-7B shows an equivalent circuit when taking into consideration the system being bonded. In figure 6-7B, the inherent inductance of an equipment case is represented by L_c, its capacitance to other equipments and the reference plane is indicated by C_c, and the impedance contributed by the insertion loss measuring circuit is represented by L_m. Measurements confirm this representation, but additionally denote that, under certain circumstances, the dip in shielding effectiveness caused by resonant effects can go below 0 dB (-10 to -20 dB have been noted), indicating that the induced voltage on the unit being bonded can be increased by attachment of a bondstrap.

Figures 6-10 and 6-11 contain plots of measured bonding effectiveness of two bond straps of different lengths that are connected between an equipment case and a ground plane. The proximity of the bond strap to the ground plane is also a parameter in the plots. The bonding effectiveness is the difference (expressed in dB) between the induced voltages on the equipment case with and without the bond strap. A negative value of bonding effectiveness in the plots indicates that the bond strap increases the amount of voltage developed on an equipment case at the indicated frequency.

It should be noted in figures 6-10 and 6-11 that about 100 MHz for 9-½" straps and above 400 MHz for the 2-3/8" straps (but below the frequencies where the length of the bond is a significant part of a wavelength), a bond will generally not affect the voltage developed on the equipment case. Also, the bonding effectiveness of the 2-3/8" straps are typically 20 dB better at 10 MHz than with the longer straps.

The data in figures 6-10 and 6-11 show a significant frequency range for each bond where the bond creates the negative bonding effectiveness condition cited earlier. This range can be increased in frequency range for each bond where the bond creates the negative bonding effectiveness condition cited earlier. This range can be increased in frequency (and also reduced somewhat in frequency range) by employing shorter bonds located away from the ground plane.

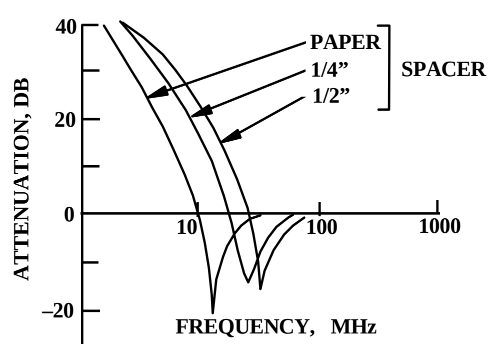


Figure 6-10. Bonding Effectiveness of 9½-inch Bonding Strap, Measured Using Shunt-T Insertion Loss Technique

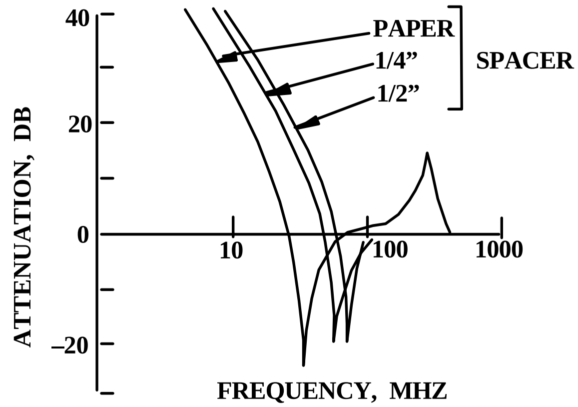


Figure 6-11. Bonding Effectiveness of 2 3/8-Inch Bonding Strap, Measured Using Shunt-T

In selecting a bond strap, the resonant frequency of the strap must be well above the highest interfering frequency which is expected to be encountered. This resonant frequency corresponds closely to the frequency at which the worst bonding effectiveness value is obtained.

To reduce the RF impedance of the bond and thus increase its bonding effectiveness, a high C to L ratio should be achieved, by minimizing the case to ground spacing and the length to width ratio of the bonding strap. The inductance of the bonding strap can be minimized by selecting a strap whose length is less than five times its width. A good bond should have an inductance of less than 0.025 microhenry .

6-3.3. BONDING RESISTANCE. Although measurement of the dc resistance of a bond is obviously not an indication of the bond’s ac characteristics, it is often used as a guide to the anticipated performance of the bond. Depending on the purpose of the bond, military specifications designate the maximum dc resistance allowable for a good bond.

For example, bonds that are installed to prevent shock hazards are required by both MIL-B-5087 INT AMD 3 and MIL-STD-1310 (SHIPS) to have a resistance of less than 0.1 ohms. RF bond requirements in MIL-B-5087 specify a resistance of less than 2.5 milliohm. Additionally, in areas prone to explosion or fire hazards, maximum values of bond resistance are designated; these values are functions of anticipated maximum fault current in the event a power line-to-ground short occurs. A guideline for a good RF bond is a dc resistance value of between 0.25 and 2.5 milliohms.

Measurements of various types of bond straps indicate the flat, solid strap provides less inherent ac resistance than other types. However, this advantage is often a minor one, and may not offset the advantages of flexibility of braid and the lower cost of solid wire.

6-4. SURFACE TREATMENT.

Both direct and indirect bonding connections require metal-to-metal contact of bare surfaces. It is frequently necessary to remove protective coatings from metals to provide a satisfactory bond. The area cleaned for bonding should be slightly larger than the area to be bonded. Ridges of paint around the periphery of the bonding area can prevent good metal-to-metal contact. Washers or fittings must fit inside the cleaned area. Immediately prior to bonding, all chips, paint, grease, or other foreign matter must be removed with a proper cleaning solution.

After bonding, the exposed areas should be refinished as soon as possible with the original finish. However, if the paint used is too thin, refinishing paint may seep under the edges of bonded components and impair the quality of the bond.

A suitable conductive coating may be used when removable components must be provided with a protective finish. Where aluminum or its alloys are used, corrosion-resistant finishes that offer low electrical resistance are available. Refer to figure 6-12 for data showing the degradation of shielding effectiveness of metal-to-metal joints caused by anodizing aluminum, and the characteristics of various conductive aluminum finishes.

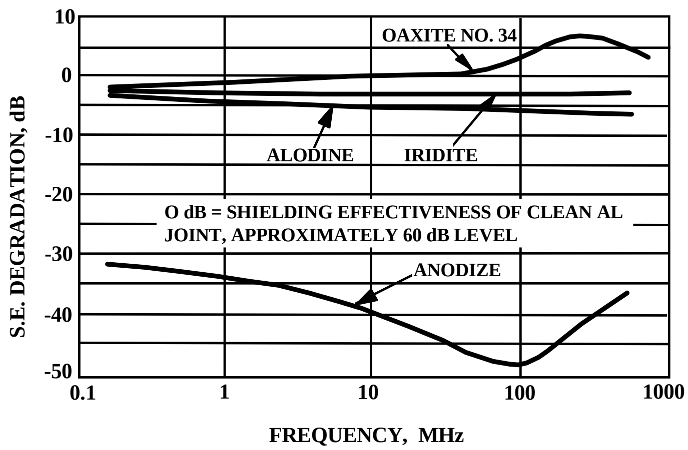


Figure 6-12. Shielding Effectiveness Degradation Caused by Surface Finishes on Aluminum

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CHAPTER 7

GROUNDING

7-1. GENERAL.

Grounding involves the establishment of an electrically conductive path between two points, with one point generally being an electrical/electronic element of a system and the other being a reference point. When the equipment element of concern is a complete ordnance system, then the reference point may be a ship or aircraft structural member. When the system element of concern is a circuit within the ordnance, then the reference point can be the equipment case or a ground plane that may or may not be isolated from the case.

A good, basic ground plane or reference is the foundation for obtaining reliable, interference-free equipment operation. An ideal ground plane would be a zero-potential, zero-impedance body that can be used as a reference for all signals in the associated circuitry, and to which any undesirable signal can be transferred for its elimination. An ideal ground plane would provide equipment with a common potential reference point anywhere in the system, so that no voltage would exist between any two points. However, because of the physical properties and characteristics of grounding materials, no ground plane is ideal, and some potential always exists between ground points in a system.

The extent to which potentials in the ground system can be minimized and ground currents reduced will determine the effectiveness of the ground system. A poor system, by enabling these spurious voltages and currents to couple into a circuit, subassembly, or equipment, can degrade the shielding effectiveness of well-shielded units, can essentially bypass the advantages of good filters, and can result in electromagnetic interference (EMI) and resultant HERO problems that may be rather difficult to resolve after-the-fact.

Various ground systems must meet requirements for personnel safety, as related to the electrical power system, for lightning protection of personnel and property, and for providing a signal ground bus as a common electronic circuit return. Emphasis in this chapter is placed on the establishment of good signal and control line grounds, since this aspect of grounding will be of most concern to ordnance design engineer.

It is important to note that the designer must consider grounding from a system point of view. The system involved includes not only the equipment or system, but all test and checkout systems. The ground systems must be compatible under the interfaced condition.

7-2. THE GROUND PLANE.

A ground plane is the mechanism by which an attempt is made to reference a number of electrical and/or electronic units to the same electrical potential so that a minimum potential exists between the interfaced units. It is not necessary for a ground plane to connect to earth ground. A ground plane may be considered as an equipotential surface or conductor and can be a flat conductive area, interconnecting wires, or tubular conductors to reference enclosures or chassis to essentially the same potential or it can be used as a separate reference for all signal circuitry. The physical configuration and the conductivity of the material will determine the intrinsic impedance of the ground plane. Other factors such as connection contact area, contact pressure, current distribution, and the overall dimension as a function of wavelength will determine the performances characteristics of the ground plane.

The dimensions of a ground plane in a system can become critical relative to the wavelengths used in the system. Hence, the ground plane must be defined. Many system descriptions never mention a ground plane, only a ground point, single reference point, vehicle ground point (VGP), etc. But, since all the system components must be connected to this ground point, the connecting network forms the ground plane.

Figure 7-1 illustrates one of several system ground planes. The system ground plane could be connected to an earth ground. Since no conductor has zero impedance, a potential exists between the system components and an antenna effect results. Hopefully, this unwanted antenna is not very efficient, otherwise transmission of precisely those signals which should leak off the ground would occur.

In order to keep this antenna inefficient, the effective length of the antenna components, the ground plane, should be kept down to a small fraction of a wavelength of the signals present. A useful rule of thumb that was empirically derived suggests that the largest effective dimension of the ground plane (L_{MAX}) be less than l/50 or, in terms of frequency in megahertz:

$$f_{MHz} < \frac{20}{L_{MAX}} \qquad (7\text{-}1)$$

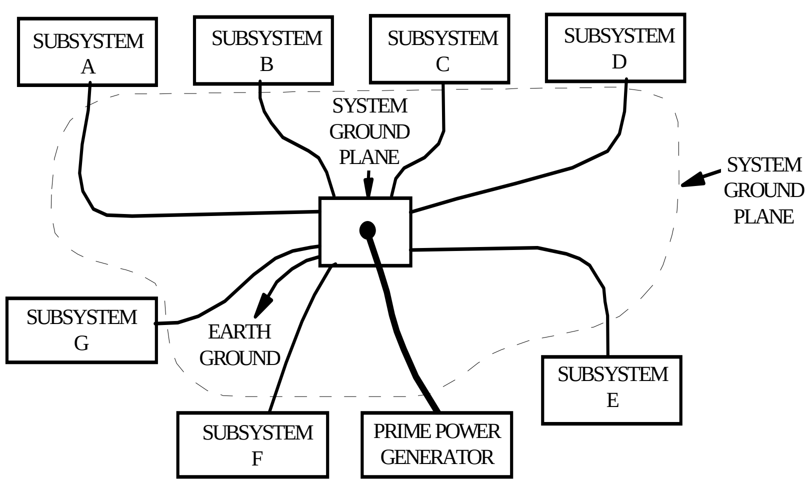

Figure 7-1. The System Ground Plane

This rule of thumb for the critical dimensions of the ground plane applies to another possible source of interference.

The electrical energy used to power electrical and/or electronic devices is usually either dc or ac of from 25 to 400-Hz, though some military equipment operates from 1000-Hz power sources. The distribution of this power by wiring, which often is neither uniformly twisted nor adequately shielded to reduce external fields, creates magnetic-induction fields. The physical configuration of the ground plane, the conductivity of the ground plane, the permeability of the conductive material, the frequency, and the flux density of the induction field will determine the magnitude of the eddy current induced into the ground plane by the varying magnetic flux. The eddy currents will generate potential differences in the ground plane and become the equivalent of a generator. This potential will couple into affected electronic devices if multiple grounding is used.

Modern systems seldom have only one ground plane. To avoid interference between various system functions, as many separate ground planes as possible are used. For instance, separate ground planes in each subsystem for structural grounds, signal grounds, shield grounds, and ac prime and secondary power ground would be desirable. Naturally, these individual ground planes from each subsystem are finally connected, by the shortest route, back to the system ground point where they form the overall system ground plane.

Grounding philosophy should be a function of unit, subsystem, or system frequency spectrum susceptibility, separation between units, subsystems, and systems, the dimensions of the ground plane used for the ground reference, the induction field ambient levels, the length of the connection to the ground plane, and the configuration of the conductor used to reference the electrical and/or electronic unit to the ground plane. The spurious generation capabilities of the electrical and/or electronic equipment will have to be considered, since multiple grounding or referencing to ground will be necessary to limit the radiated field of spurious energy.

7-3. GROUNDING TECHNIQUES.

There are three fundamental grounding methods that can be employed, as illustrated in figure 7-2. The approaches can be used separately or in combination in any given equipment or system. They are Floating Ground System, Single-Point Grounding System, and Multi-Point Grounding System.

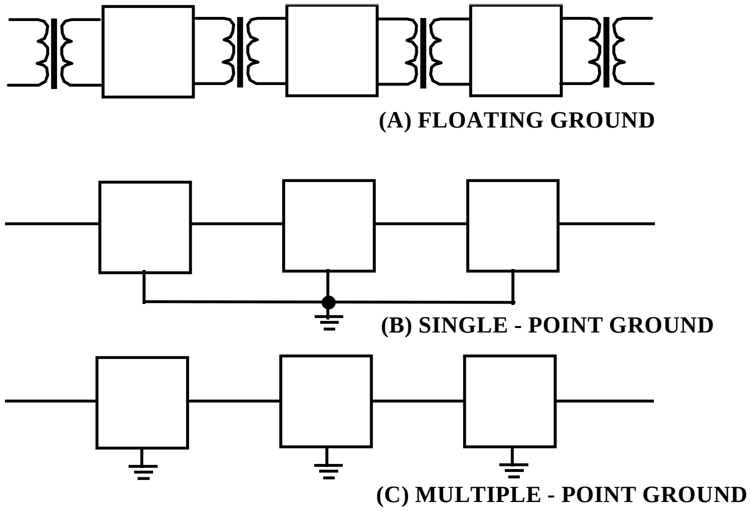

Figure 7-2. Grounding Methods

7-3.1. FLOATING GROUND. The Floating Ground System is a method to electrically isolate circuits or equipments from a common ground plane, or from any common wiring that might introduce circulating currents. Floating grounds depend, for their effectiveness, on truly "floating."

For many situations, this complete isolation may be very difficult to achieve. This is particularly true of ordnance that will be hard-wired to an aircraft or launcher. Also, certain hazards exist in the use of floating systems, in that static charges or lightning potentials may accumulate between the floating grounds and other accessible grounds such as the flight-deck, weather-deck or other portions of the ship structure, power line neutrals, the skin of the weapon system, etc.

Examples of isolation techniques are shown in figures 7-3 and 7-4. Figure 7-3 employs transformer isolation, while figure 7-4 obtains isolation by optical means.

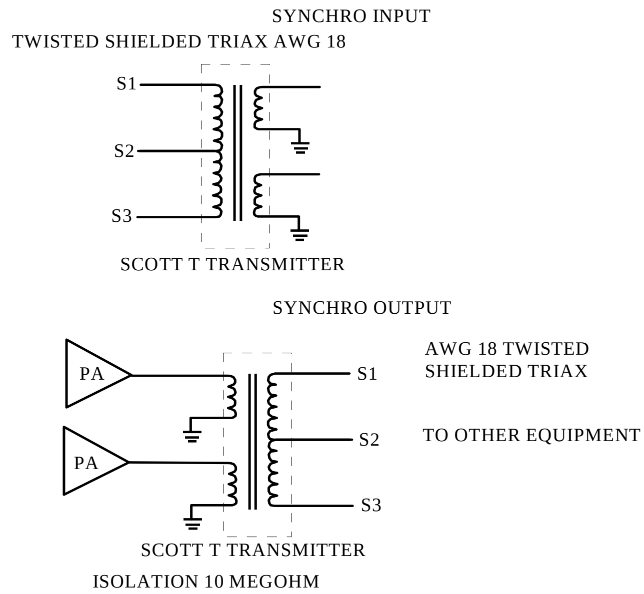

Figure 7-3. Typical Synchro Input and Synchro Output Isolation Circuitry

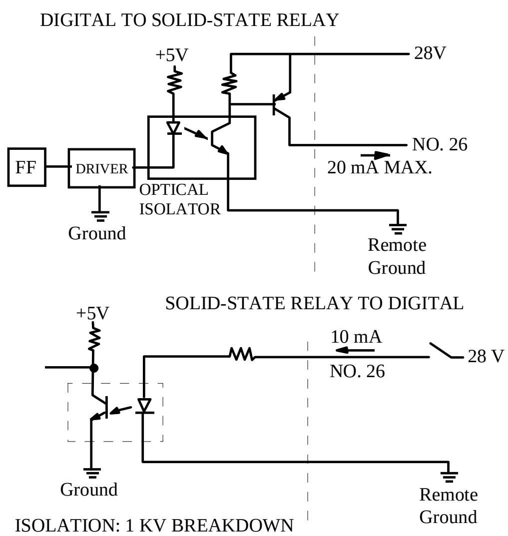

Figure 7-4. Digital To Solid-State Relay, and Solid-State Relay to Digital Isolation Circuitry

7-3.2. SINGLE-POINT GROUND. In a Single-Point Grounding System, a single physical point in the circuitry is defined as a ground reference point. All ground connections are tied to this point.

For multiple enclosure configurations, the enclosure and electronic circuit grounds are often kept separate, with the single-point grounding concept being used independently for each ground system. The interconnection between ground systems is then made only at the reference point. This isolates the enclosures, and prevents circulating currents in one ground system from affecting the other.

Again, because of the interconnections of control, test and basic ordnance units, it would be extremely difficult to maintain a complete single-point grounding system. However, under certain conditions, it may be advantageous to use single-point grounding for the basic ordnance and combination grounding for the total weapon system.

At high frequencies (frequencies whose wavelengths approach equipment ground plane dimensions or cable lengths), single-point grounding systems are no longer practical. Therefore, it is important for the ordnance equipment/system designer to have an understanding of the frequency-susceptibility of his device, as well as the internal and environmental frequencies that are around to create EMI, before finalizing the grounding system design.

7-3.3. MULTIPLE-POINT GROUND. The Multiple-Point Grounding System is one in which many connections to the ground plane are used instead of individual return wires for each circuit. The ground plane might be an equipment chassis, or a ground wire that is carried throughout a system.

The advantages of a multiple-point grounding approach are that circuit construction is easier, and it is the only way to avoid standing-wave effects in the ground system at high frequencies. Also, a large conductive mass can be selected for the ground plane. However, since multiple-point grounding creates many ground loops, the quality of the ground system becomes very important. Also, the ground plane must be carefully maintained, particularly with reference to corrosion, vibration, and mechanical damage, which can introduce high impedances into the ground system.

7-4. THE GROUNDING SYSTEM DESIGN OBJECTIVES.

To accomplish interference suppression through an effective grounding system, all the system electrical and structural components must be maintained at the same reference potential. This is generally accomplished by setting up separate grounding systems for these separate grounds grounds only at one common reference point or plane. The object is to prevent any EMI generated by one unit of the system from being transferred through a common ground impedance to the other units. If potential differences are not allowed to exist, interference currents cannot flow and spurious signals can neither be radiated or conducted to the susceptible parts of the system. Obviously, the larger the system, the less this ideal can be achieved. No conductor has zero impedance, hence zero potential difference can only be approached, never completely achieved.

In order to minimize the potential differences between the parts of the system, the various equipments are grouped by their characteristics, and each group has its own separate ground system as indicated in the following paragraphs.

7-4.1. STATIC AND STRUCTURAL GROUND. These grounds take in all those conductive parts of the system that are designed not to carry current. These parts include mechanical strength members and mechanical parts and enclosures that contain low-frequency circuits such as static shields, cable shields, chains, and black boxes," including junction boxes. Low frequency is defined in megahertz as:

$$f_{MHz} < \frac{20}{L_{MAX}} \tag{7-2}$$

Where L_{MAX} is the maximum dimension of the part in feet, including its connection to the system ground plane. Each of these items is grounded to the structural ground at only one point using proper bonds. It is important that no electrical circuitry be connected to the static or structural ground system.

Any connections of static or low-frequency cable shields to the ground system should be routed through the pins of any connectors encountered. If a cable shield appears between two static shields (for the low-frequency case), it generally accepted that the cable shield be connected to one of the static shields, but not to both. The other end of the cable is left floating. The combined shields and the remaining static shield each are connected separately to the structural ground.

Shields enclosing high-frequency circuits and cables must be designed to ensure that all openings, covers, and connectors are RF tight. High-frequency circuits are defined as circuits containing a maximum frequency (in megahertz) of:

$$f_{MHz} > \frac{20}{L_{MAX}} \tag{7-3}$$

High-frequency shields are grounded to the static ground at least at both ends and also at intervals, depending on the frequencies being shielded.

When both high and low frequencies are present, the designer is confronted with a dilemma, should be single-point or multiple-point ground? A compromise can be reached by connecting one end directly to the ground system and the other end through an appropriate capacitor, according to the high and low frequencies present, to the static ground system, using the relationship:

$$X_{cl} = \frac{f_h}{f_L} X_{ch} \tag{7-4}$$

where

f_h = the lowest of the high frequencies that requires multiple-point grounding.

f_L = the highest of the low frequencies that requires single-point grounding.

X_{ch} = the impedance between shield and ground that f_h can see to still be effectively multiple-point grounded.

X_{cl} = the impedance between shield and ground that f_L that must see to preserve single-point grounding at that frequency.

7-4.2. AC PRIME POWER GROUND. This is the power obtained from the host vehicle or ship. Under conditions using power generator, the National Electric Code (NEC) requires that before entering the user building, the neutral must be earth-grounded. Within the building and within the equipment, the primary power ground is left floating. Aboard a vehicle or a ship, prime power is generated and the neutral is referenced to the vehicle, or ship ground plane or ground point.

Any secondary ac power that is derived, but separated from the primary ac power, has its own ground system with the ground returned to the secondary power source. At each power source, the secondary power ground system is then grounded to the system ground plane.

7-4.3. DC POWER GROUND, SIGNAL GROUND, OR CIRCUIT GROUND. DC power is supplied to circuitry using frequencies from dc up into the gigahertz range. The dc ground is often the most versatile and complicated of the system grounds. There are a number of special cases where a dc supply cannot be grounded at all, but in general, the zero-volt dc return of every dc power supply should be connected separately at one point to the system ground point.

Seldom can the ideal grounding combinations be achieved in practice, but the designer should strive for the most advantageous combination. With this in mind, the following ground rules are presented:

a. Supply low-frequency and high-frequency circuitry from separate dc supplies.

b. Supply low-signal-level and high-signal-level circuitry from separate dc supplies.

c. The zero-volt terminal of the dc supply is the dc power ground.

d. Connect the dc ground to the dc power ground of its own supply, and only of its own supply.

e. Connect the dc power ground at one point to the system ground point or ground plane.

f. Where several power supplies of different dc potentials must share a common dc power ground, interconnect the zero-volt terminals of each supply and make a single-point connection to the system ground point.

g. Run separate grounds to the dc power ground from each module or critical stage. In some cases, several noncritical modules containing functionally similar circuitry can have a common ground lead to dc power ground. Engineering judgement must determine the degree of application of these rules.

h. For RF circuits, the circuit parts and interconnection wiring are often mounted on (and close to) a ground plane formed by a copper sheet and grounded directly to the system ground plane.

i. When coaxial cable is used for signal transmission in high-frequency circuitry:

$$f_{MHz} > \frac{20}{L_{MAX}} \qquad (7\text{-}5)$$

the outer conductor is connected to the circuit grounds at both ends. If both circuits are also separately grounded to their dc power grounds or returns as shown in figure 7-5, a ground loop is formed into which low-frequency current can be induced by magnetic flux linkages from and outside source. In many cases, this may not affect the high-frequency circuitry, but the loop can transmit the low-frequency energy to another susceptible circuit. As a first step, try to open the loop. One of the ground returns might not be necessary. Or, try to reduce the loop area by running ground leads close together and shielding them. In connection with this, a triax has an advantage over the coax by allowing the use of the outer conductor as a shield that is multiple-point-grounded to the system ground plane, thereby forming a shielded tunnel for the coax. But, the induced current flowing around the ground loop depends on the magnetic flux linking the loop.

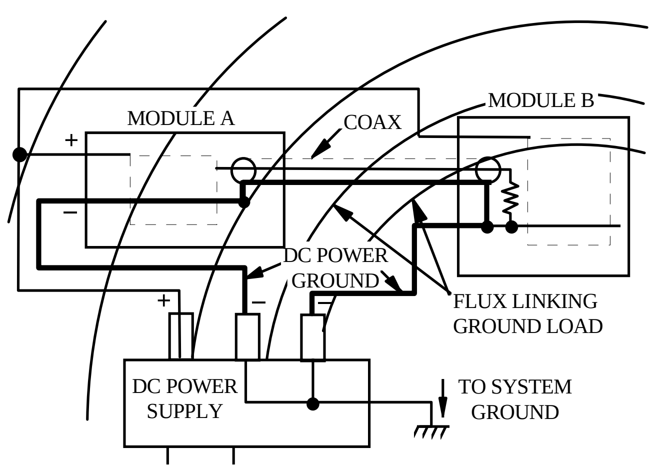

Figure 7-5. Flux Linking a Ground Loop

An often effective method to reduce the flow of circular current in a ground loop is the isolation of one or more of the circuit ground connections by inserting a resistance between the circuit ground and its dc power ground connection. If some signal gain can be spared, a resistive coupling can at times be inserted between the coax outer conductive and ground. In certain cases, ac coupling, especially frequency-selective ac coupling, has been used with success.

j. It is common practice to manufacture standard cable assemblies containing multiple conductors. The spare conductors in these cable assemblies should be single-point-grounded with half of the spare conductors at one end and the remaining half of the spare conductors at the opposite end. These spare conductors, when grounded in this manner, will function as effective electrostatic (Faraday) shields within the cable assembly, thereby providing isolation between adjacent signal circuit conductors and reducing cross-talk due to capacitive coupling.

7-4.4. SHIELD GROUNDING. The method of terminating a shield to a ground plane will determine the shielding effectiveness in accordance with frequency. Since, as frequency increases, current flow is either closer to the surface or at the surface depending in the configuration of the conductor, then peripheral grounding of a shield will produce the most effective connection of a shield. The braid which commonly serves as a cable shield acts as an effective barrier to low-frequency electric fields, but at shorter wavelengths, when the electric and magnetic components of the wave front become equal in magnitude, it also becomes effective for the electromagnetic field. The distributed capacitance existing between the conductor and the shield will couple some energy to the shield. The shield, in absorbing the radio frequency energy, will develop an induced current. When the shield is connected to a ground plane through a minimum impedance, no appreciable potential can develop and of itself will not serve as a radiator of radio frequency energy. Figure 7-6 shows the effects of the lead length of the conductor used to connect a cable shield to a ground plane. The tests were performed starting at 40 MHz to realize an electromagnetic field.

A coaxial cable, though multiple grounded to the dc power or signal ground, will still have the return signal current flowing through the shield. However, it is sometimes inconvenient to ground a shield often enough to restrict radiation from a coaxial cable shield. In such cases, a double-shielded cable, such as triax, where the two shields are electrically independent of each other will provide the additional attenuation required for electromagnetic compatibility. In such cases to fully realize the shielding capabilities of both shields, the shields should be independently terminated to the ground plane and not through a common connection. By independent connection of each shield to the ground plane, up to a 30-dB improvement can be obtained. An additional 10- to 12-dB improvement can be realized by independently terminating each shield to independent ground plane, i.e., the inner shield to the signal ground and the outer shield to the static ground. Connectors are available which allow independent peripheral grounding of each shield, thus affording maximum EMI protection at radio frequencies.

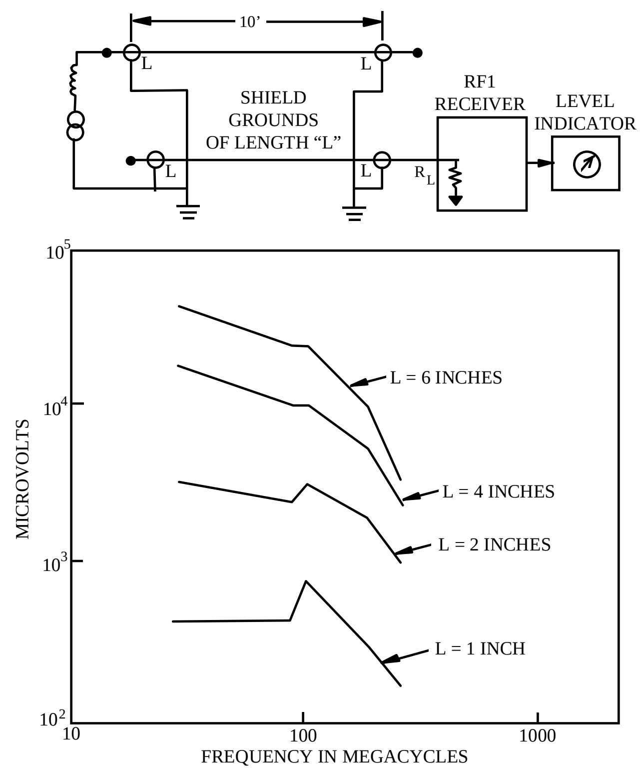

Figure 7-6. Coupling as a Function of Termination Length of Shield

7-5. CORROSION OF GROUNDING MATERIALS.

Electronic and/or weapon system equipment mounted or enclosed by metals can oxidize readily when in contact with other metals. These effects bear consideration when providing connections to ground such metals. For example, aluminum will corrode and, unless suitably protected, will oxidize very rapidly. Aluminum oxide is a nonconductor and thus a contact area containing aluminum oxide will act as an antenna and the oxide can act as a nonlinear rectifier junction, in turn creating additional frequencies which then can be reradiated by the connecting wire. To reduce the oxidation rate for aluminum, semiconductive coatings such as iridite or allodize are applied to the aluminum surfaces. These coatings are nominally very thin, in the order of micro-inches. If a metal such as copper is used as the connecting material between the aluminum and ground, an electromotive force potential of 2.044 volts will exist between the aluminum and copper. An equivalent battery potential of 1.655 volts will exist between aluminum and iron/steel.

Magnesium displays higher electromotive force potentials of 2.744 volts between magnesium and copper and 2.355 volts between magnesium and iron/steel. Iron/steel, of course, has a lower order of conductivity (approximately seven percent) when compared to copper. The point here is that for effective grounding, which depends on good dc and ac connections, the galvanic actions, the electromotive force valence potentials, the oxidation rate, and mating materials must be considered, otherwise additional spurious frequencies will be generated, and the equipment will, over a short period of time, assume a potential other than ground and operate as an antenna to either receive energy or transmit energy, resulting in equipment malfunction.

Material consideration should also involve migration of metals such as silver, zinc (galvanized iron), and aluminum. Silver plating of conductors or buses should be held to a minimum, particularly when other conducting materials are in close proximity, to reduce the eventual possibility of creating short circuits. Copper, being a more stable metal, can be obtained with a minimum conductivity of 101 percent (electrolytic tough pitch copper = 100) and thereby achieve the low resistance to current flow as displayed by silver. Migration of zinc and aluminum form whiskers. Again, short circuits can result if conductive materials are in close proximity.

To retard oxidation, iron or steel are frequently cadmium plated. Cadmium has a valence potential of 0.401 volt relative to hydrogen. However, cadmium has a low order of conductivity, approximately 22.7 percent compared to copper. To achieve a low-resistance connection, such as 0.0005 ohm or less, a sufficiently large contact area is required as well as a minimum tension requirement. When conductive material of low orders of conductivity has to be used, the total contact area has to be increased. Since it is difficult to instrument and therefore determine the contact resistance at radio frequencies, contact resistance should be specified as a dc resistance and measured by a double Kelvin bridge. If proper selection of mating conductive materials has been made in regard to corrosion, electrolysis, and material shape or configuration, the ground connection at radio frequencies should not have a much higher value than at dc.

7-6. CIRCUIT GROUNDING CONSIDERATIONS.

If the ground plane is fairly large, significant potential differences may exist between two points on the ground plane. These voltages must be considered when defining the permissible ambient interference level in the system.

The simplest and most direct approach to keeping potential differences introduced by the circuit ground plane to a minimum is to arrange circuit components physically so that ground return paths are short and direct, and have the fewest possible crossings of these paths. In this way, the inter-circuit coupling of the ground currents will be low and isolated.

The effect of ground potential can be cancelled by electrically isolating the circuits using a floating ground system. This method is especially effective at audio and low radio frequencies. Above these frequencies, however, its effectiveness progressively diminishes because, as the frequency of equipment operation increases, coupling paths appear that bypass the isolation transformers.

Band pass filters are another method of obtaining circuit isolation. These filters are only effective when the frequency of the interference energy is outside of the frequency band of the desired signal. Then the band pass filter will reject the interference signal across the load, causing the voltage drop to occur elsewhere in the circuit and not at the load. When using band pass filters, care must be taken to ensure that "ringing" is not initiated by the undesired signal when it is processed by the filter.

Differential or balanced circuitry can help reduce the effects of ground circuit interference. Since a differential circuit responds only to the potential difference between its input leads, the noise voltage at the source may be above ground potential by a considerable amount without degrading circuit performance. Figure 7-7 illustrates such a differential circuit. V_g is the voltage to which the device responds. V_n, the interference voltage, is simultaneously impressed on both input leads but is balanced out in the input to the device because each input lead has the same impedance to ground. Thus, the device does not respond to the ground circuit signal. In theory, the ambient noise voltage is cancelled out, assuming the impedance of V_g is zero ohms. In practice, there is always some unbalance in the differential device or associated circuitry and some part of V_n will appear as a difference voltage across an equivalent resistance R. The noise voltage differential results in a reduced output signal-to-noise ratio. Figure 7-8 shows how the unbalance causes a portion of V_n ΔV_n at the input to the differential circuit.

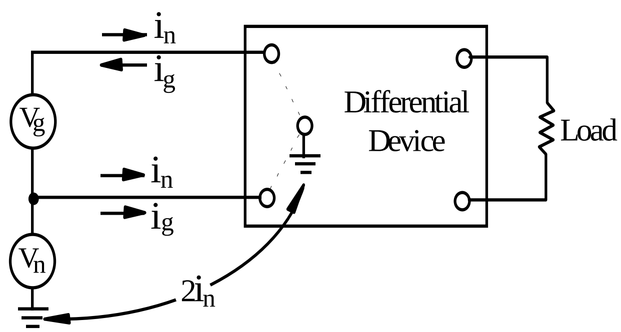

Figure 7-7. Schematic Diagram of a Differential or Balanced Circuit

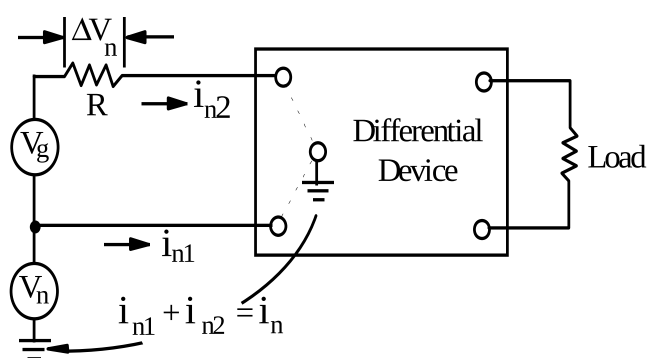

Figure 7-8. Effect of Unbalance in a Differential Circuit

Figure 7-9 summarizes the above discussion. It shows the ground circuit voltage, V_n, introducing an incremental voltage, ΔV_n, at the input to the differential circuit.

In digital circuits, the input and output impedance of the circuits are generally relatively low. This makes the circuits more susceptible to the effects of low impedance (magnetic) fields than to the effects of high impedance (electric) fields. One of the important parameters controlling interference due to a low impedance field is the loop area of the circuits causing and picking up the interference. By minimizing the loop area of these circuits, much of the interference problem can be eliminated.

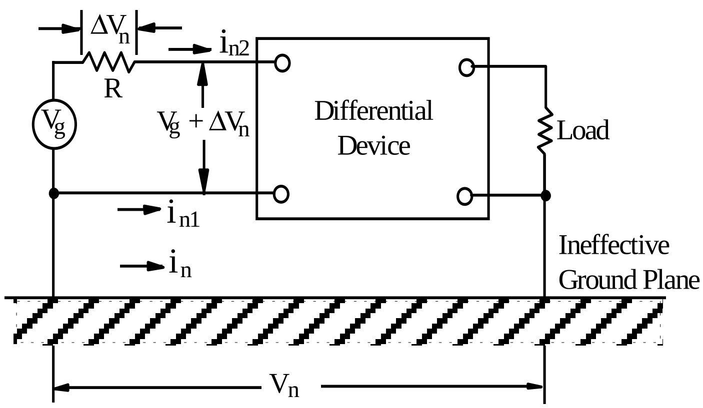

Figure 7-9. Common-Mode Voltage Generated by Current Flowing in a Finitely Conducting Ground Plane

Where modular-type construction is used, one method of minimizing the loop area of grounded circuits is to mount modules on a sheet of good conducting material, such as copper or aluminum, that is connected to the circuit ground as directly as possible. All intra-module wiring should be run as close to this sheet as possible. This technique can reduce the loop area of these circuits to an extremely small value, but the capacitive coupling of the circuits is increased. The effect of this increase in capacitive coupling on circuit operation should be considered if this technique is used.

The use of high impedance input or output circuit impedances when signals must be transmitted over even a few inches should, in general, be scrupulously avoided. Where the use of such signals cannot be avoided, the interconnecting lead must be shielded and the shield grounded at each end.

7-7. POWER SUPPLY CONSIDERATIONS.

It is generally good practice to isolate power and signal grounds from each other. This will go a long way to minimizing the possible coupling of signals between these two major types of paths.

Groups of electronic circuits can be connected to one power supply based on a pre-arranged logical pattern. Critical circuits must be kept separated (i.e., critical with respect to interference susceptibility), and each given their own power supply when possible. If it is decided that these critical circuits cannot be kept separate, a voltage regulator can be introduced to act as decoupling component; Zener diodes can usually be used quite effectively in such applications.

Use of a large capacitor or an RC network at the load end of power supply leads can offer some degree of decoupling. This is particularly helpful in digital circuits to reduce the inductive effects of long power supply leads, but can introduce "ringing" at frequencies where the line and capacitor resonate. This condition can be circumvented by shunting the decoupling capacitor with a small RF bypass capacitor.

When deciding on which circuits to connect to which power supplies, the physical location of the electronic circuits must be considered. If no other factors are involved, all circuits fed by a given power supply should be grouped in the same area. However, in cases where various circuits will have to operate simultaneously (as in a synchronous digital system) some spatial separation may be advantageous to prevent coupling of radiated energy.

Alternating currents of high amperage are potential sources of interference coupling to adjacent lines. Transposed or twisted lines should be used for all ac power circuits. Routing of these lines should be away from susceptible lines. AC power circuits in which switching transients are expected may beneficially use shielding to contain the higher frequency components of the transient. Such shielding should be over the entire power wiring bundle and should be grounded at both ends.

A power system that drives electro-mechanical relays should not generally be used as a supply voltage for any electronic circuits because of the large amount of noise in this system due to relay operations. In addition to providing power relays, this power source may also furnish power for indicator lights, small motors, electromagnetic clutches and brakes, and these devices may also be sources of noise. The power supply itself generally consists of a transformer-rectifier combination, or a motor generator set. In either case, the source is usually regulated to ensure a constant voltage over a wide range of loads.

The primary cause of relay interference is due to the de-energizing of relays. When a relay circuit is opened, the collapsing magnetic field of the relay coil induces a large voltage of the opposite polarity across the coil. This induced voltage initiates an arc across the contacts that are breaking the circuit. The arc contains energy, in a wide band of frequencies, that may be transmitted as interference both by conduction and radiation. There are many standard arc suppression techniques, but even when these are used, a certain amount of noise will be conducted or radiated from the source.

Another cause of conducted interference can be the changes of current in the relay power supply busses due to the continually changing demands when the equipment is in normal operation. The conducted interference can be reduced by isolation of the relay circuits from the rest of the equipment.

To maintain isolation of the relay circuits and to prevent the system from floating and possibly reaching a high potential above ground, the relay system should preferably be grounded at a single point. Because of the unusually wide distribution of relay circuits, and because it is the point of maximum current, the ground connection should be made at the source.

In designing the power distribution system for relay systems where large currents are expected, it is important to consider the IR drops of the conductors in addition to their current-carrying capacities, although there is relatively low voltage involved. Also, the generation of varying magnetic fields due to the variation of the relay power supply currents must be considered. The effects of these variations can be minimized by running the supply and return leads as close together as possible throughout the system, thereby reducing the loop areas of the various relay circuits.

To minimize the possibility of dc supply lines coupling interference to other circuits, and to protect them from receiving interference from other lines, the best procedure is to use a twisted-pair for the supply line and its return. Shielding offers little advantage on lines of this low impedance. The routing should be away from ac power and control lines. When possible, run separate supply lines from the various circuits using the supply to the supply output. Do not share supply return lines with other circuit returns.

7-8. PRIME POWER CONSIDERATIONS.

There are two primary concepts regarding power returns. These are ground or structure common return and wired return. Although the ordnance designer will probably have no control over the prime power configuration into which his equipment must work, he should nevertheless be familiar with these two concepts.

In the ground or structure common concept, one side of the power system is grounded at the power source, and all loads use the vehicle frame or structure as the return conductor. In three-phase connected ac systems, the neutral is grounded and all single phase loads use the structure for the return circuit. The principal advantage of this system is the reduced weight resulting from the elimination of a great many heavy power return wires.

The disadvantage of the common return system is that it does not distribute the power efficiently. The flow of currents through structure produces voltage drops in the structure. These voltages are normally small compared to the operating level of the power system, but are large compared to the operating level of electronic systems. This can create potential interference problems in any electronic system using the structure as a power return. Even systems using structure as a ground plane for shield grounds are subject to induced voltages in susceptible circuits.

In the wired return concept, all systems are grounded at one point only and have wires for all return circuits. The wired return system eliminates the vehicle structure as an impedance common to all systems and therefore eliminates the complex ground loops which exist with the common return system.

7-9. CABLING CONSIDERATIONS.

The problem of EMC in a complex electrical or electronic system is, in many cases, dependent on the treatment of the shielding and the grounding of the shields of interconnecting leads. Injudicious application of a grounded shield to a wire may cause coupling problems that otherwise would not exist.

Grounding of the shields may be accomplished as single-point or multiple-point grounding. Factors that influence the selection of single-point or multiple-point grounding include the interference signal frequencies involved, the length of the transmission line, and the relative sensitivity of the circuit to high- or low-impedance fields.

7-9.1. SINGLE-POINT SHIELD GROUNDING. For multilead systems, each shield may be grounded at a different physical point as long as individual shields are insulated from each other. Single-point shield grounding is more effective than multiple-point shield grounding only for short shield length. Single-point grounding is ineffective in reducing magnetic or electrostatic coupling when conductor-length-to-wavelength (L/λ) ratios are greater than 0.15, where the wavelength is that of the highest frequency to be used (or the highest frequency interference to be expected) on the wire or in the system.

7-9.2. MULTIPLE-POINT SHIELD GROUNDING. For L/λ ratios greater than 0.15, multiple-point grounding at intervals of 0.15λ is recommended, since the shield can act as an antenna that is relatively efficient at $L = \lambda/4$ when one end is grounded. When such grounding of the shield at intervals of 0.15λ is impracticable, shields should at least be grounded at each end. Multiple-point shield grounding is effective in reducing all types of electrostatic coupling, so long as large ground currents do not exist. In general, multiple-point shield grounding solves most problems, but in audio circuits, single-point shield grounding may be more effective because of the ground current problem.

An interesting comparison of the magnetic interference susceptibility of cable-connected circuitry is shown in figure 7-10. The evaluation was performed at 100 kHz, with the measurement parameters having to do with the type of cable used, whether or not the load is grounded, and how the cable shield is grounded. Among other results, the comparison shows (for the cases indicated) the disadvantage of returning the load to the ground plane, and the advantages gained using tightly twisted leads.

The performance of tests indicate that, for low-level signals and low-impedance circuits where the distance between input connector and circuit input is small, i.e., an inch or two, the use of a twisted-pair alone may prove adequate. For long runs, the use of twisted-pair shielded wire becomes mandatory; this is true with unbalanced as well as with balanced output devices. In the balanced case, there should be no grounding except for the shield; in the unbalanced case, it is often necessary to ground one of the conductors as well as the shield. The shield and conductor should be grounded at the same point. Single-point grounding of the shield should be used for short runs and multiple-point grounding for long runs. If a shielded twisted-pair is part of a cable bundle and the bundle must go through a connector, three pins should be made available on the connector to provide insulated passage of the twisted-pair leads and the shield when it is not desirable to ground the shield at the connector.

Signals of high level will, in general, not be bothered by susceptibility of circuits. Rather, it may be a source of interference to lower level signal lines. For this reason, and dependent on other characteristics of the signal, either a twisted-pair or a shielded lead should be used. Multiple shielding may be required if the signal is of sufficient level. Grounding should be applied at both ends to prevent electric field radiation from the cable.

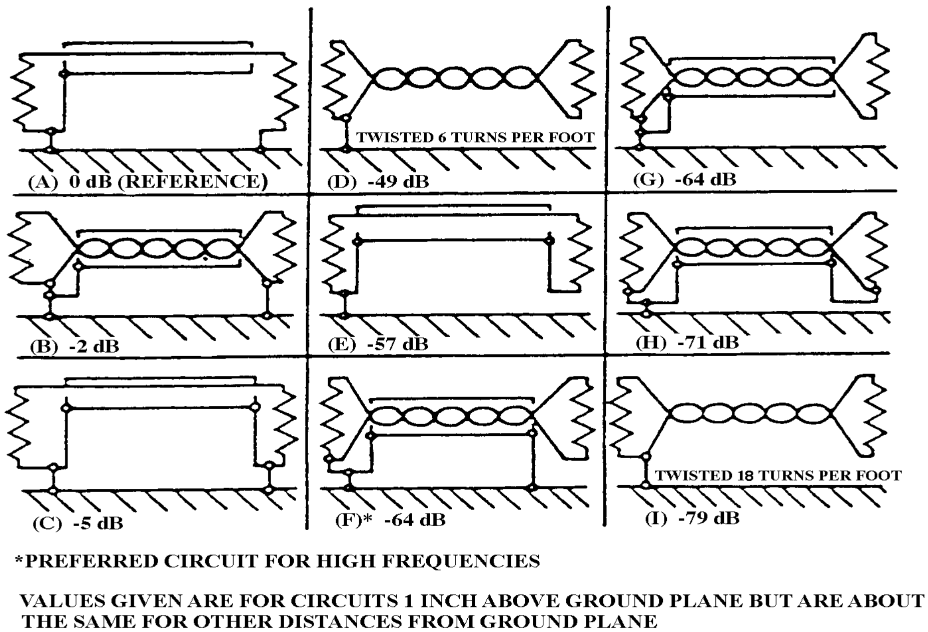

Figure 7-10. Relative Susceptibility of Circuits to Magnetic Interference

When cable shields are grounded, good electrical contact to the shield must be established. If possible, the shield should be grounded completely around the periphery of the connector shell. The use of pigtail grounding should be avoided on all cables, particularly those carrying or exposed to signals above 1 MHz. When it becomes necessary to use a pigtail, the length should be minimized.

Figure 7-11 provides an indication of how the length of these ground leads can affect the coupling between two shielded cables.

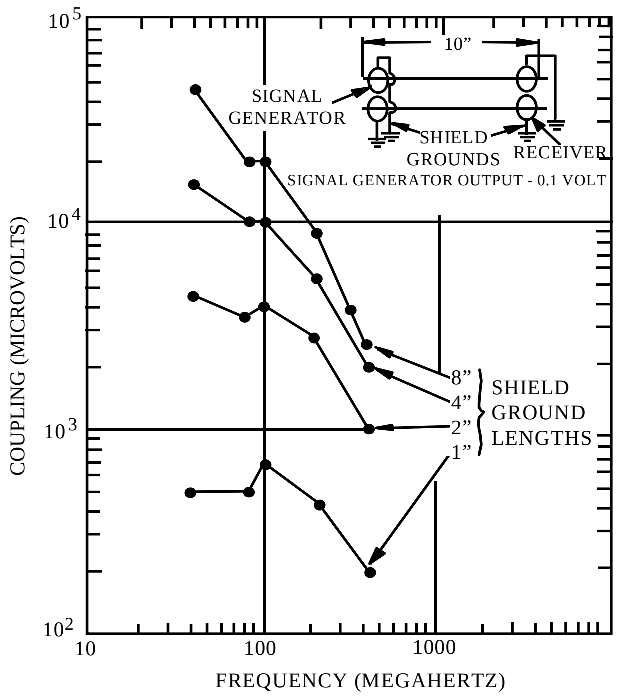

Figure 7-11. Cable Coupling Vs. Shield Ground Length

There may be situations when multiple-shielded cable usage is considered necessary. For example, one suggestion has been made to use multiple-shielded cable for unbalanced input circuits with sensitivity thresholds better than 50 millivolts and operating frequencies between 5 kHz and 1 MHz. When biaxial or triaxial cables are employed, single-point grounding should be used. An illustration of how to ground double-shielded cable is provided in figure 7-12.

Double or triple-shielded cable may be necessary for feeding high input or output impedance circuitry, particularly if the circuit is in a high EME.

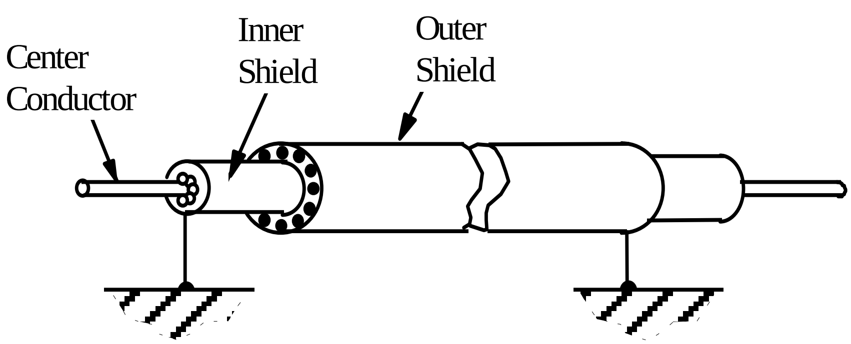

Figure 7-12. Example of Grounding a Double-Shielded Coaxial Cable

7-10. GROUNDING DESIGN GUIDELINES.

Although the specific grounding philosophy to be employed in ordnance design is very dependent on some of the detailed design objectives of that equipment, it is nevertheless possible to establish general grounding guidelines to follow. These are stated below, with the understanding that they should not be applied rigidly, and that alternate grounding methods should be tested before selecting a final design.

a. Remember that the ground system of concern includes the combined effects of both the basic ordnance and all control, operation, test and checkout subsystems.

b. Use single-point grounding when the dimensions of the circuit or component under consideration are small compared to the wavelength of concern (typically less than 0.15λ). Use multiple-point grounding when circuit or component dimensions exceed 0.15λ. When possible, ground large circuits or components at several locations, so that the separation between grounds is never greater than 0.15λ.

c. There are occasions when transformer isolation and other isolation techniques can be used to prevent ground loops from occurring.

d. Keep all ground leads as short and direct as possible. Avoid pigtails when terminating cables.

e. It is advisable to maintain separate ground systems for signal returns, signal shield returns, power system returns, and chassis or case ground. They can be tied together at a single ground reference point.

f. Ground reference planes should be designed so that they have high electrical conductivity, and so that they can be easily maintained to retain good conductivity.

g. Circuits that produce large, abrupt current variations should have a separate grounding system, or should be provided with a separate return lead to the single-point ground. This will reduce transient pickup in other circuits.

h. Grounds for low-level signals should be isolated from all other grounds.

i. Never run supply and return leads separately, or in separate shields. A twisted pair is the best configuration for the supply bus and its return. Also, avoid carrying signal and power leads in the same bundle or in close proximity to one another. When signal and power leads must cross, make the crossing so that the wires are at right angles to each other.

j. Use of differential or balanced circuitry can significantly reduce the effects of ground circuit interference.

k. For circuits that operate below 1 MHz, tightly twisted pairs of wires (either shielded or unshielded, depending on application) that are single-point grounded offer the best approach to reduce equipment susceptibility.

l. When coaxial cable is necessary for signal transmission, it appears that signal return through the shield and single-point grounding at the generator end offers certain advantages at the lower frequencies. However, other grounding arrangements should be considered. At high frequencies, multiple-point shield grounding is required.

m. Low-level sensitive transmission lines may require multiple shields. Single-point grounding of each shield is recommended.

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CHAPTER 8

EMI FILTERING SUPPRESSION DEVICES

8-1. GENERAL.

Ordnance cannot always be protected from the electromagnetic environment (EME) by shielding and circuit design alone. Firing circuits and other lines that penetrate the radio frequency (RF) shield of the ordnance can conduct electromagnetic energy to the electrically initiated devices (EID's). To protect the ordnance from the effects of HERO, these circuits must be filtered or surge-protected at their point of entry into the shielded enclosure. Electromagnetic interference (EMI) filters and a variety of transient suppression devices have been developed for this application. A list of useful suppression devices available to the ordnance designer is shown in table 8-1.

Table 8-1. Types of Suppression Devices

INDUCTANCE-CAPACITANCE-RESISTANCE (LCR) FILTERS Types: (L, T, and PI filters)
FEED-THROUGH CAPACITORS Panel and filter-pin connectors
BROAD-BAND ABSORBERS Lossy materials (Iron, Ferrites, Hybrids)
SOLID-STATE DEVICES [(Metal Oxide Varistors (MOVs), Avalanche Diodes, Gas Discharge Tubes, Thyristors]

The purpose of this chapter is to provide the ordnance design engineer with general EMI suppression design information. EMI filters, their functions and design consideration are described first, then transient suppression devices and their applications are presented.

8-2. EMI FILTERS.

While filters are necessary and should be placed where needed, care should be taken to avoid redundant filtering caused by uncoordinated efforts of separate design groups. Redundancy usually occurs when each "black box" is required to meet an interference control specification, regardless of final location. Although tradeoffs must be made, there is no substitute for a well thought-out system electromagnetic compatibility (EMC) control plan. If formulated well ahead of the system design, filter duplication can be avoided.

The effectiveness of any EMI filter is greatly influenced by the impedance of its source and load terminations. Manufacturers of EMI suppression filters normally specify the filter characteristics with fixed source and load impedances (usually 50 ohms). The actual circuit characteristics may be very different values. This aspect must be taken into consideration when designing, specifying, or using EMI filters. Refer to MIL-STD-1377 for testing filters used in ordnance.

The basic characteristic used to describe filter performance is its insertion loss. Insertion loss is defined as the ratio of voltages appearing across the system terminals immediately beyond the point of insertion of a filter, before and after insertion. This ratio is expressed in decibels (dB) as follows:

$$insertion\ loss = 20\log[E_1/E_2] \qquad (8\text{-}1)$$

where:

E_1 = load voltage without filter

E_2 = load voltage with filter inserted

A number of other filter characteristics are important. The input and output impedance requirements have already been cited. Others include the attenuation in the pass-band, the skirt falloff characteristic (rate at which the filter insertion loss changes as a function of frequency), steady-state and transient voltage ratings, etc. These parameters are discussed in more detail in paragraph 8-6.

8-3. FILTER DESIGN.

NOTE

Except for ferrite filters, only lumped constant filters are considered in this design guide.

As indicated previously, filters are designed to attenuate at certain frequencies while permitting energy at other frequencies to pass unchanged.

Reflective filters do this by using combinations of capacitances and inductances to set up a high series impedance or a low shunt impedance for the interfering currents. Lossy filters do this by absorbing the interference energy.

The passband of a filter is the frequency range in which there is little or no attenuation. The stop band is the frequency range in which attenuation is desired. The attenuation may vary in the stop band and is usually least near the cutoff frequency (the frequency at which a 3-dB insertion loss is obtained), rising to high values of attenuation at frequencies considerably removed from the cutoff frequency.

8-3.1. LOW-PASS FILTERS. EMI control usually requires filters of the low-pass type. Filters can be grossly classified according to the relative positions of the passband and stopband in the frequency spectrum. There are four classes: low-pass, high-pass, bandpass, and band-reject. The following discussions deal with low-pass filters as the most commonly used to reduce the effects of HERO. Detailed usage of all types of filters can be found in NAVAIR AD 1115. Power line filters are low-pass filters that pass dc or power frequency currents without significant power loss, while attenuating signals above these frequencies. Filters incorporated in amplifier circuits and transmitter output circuits are usually of the low-pass type so that the fundamental signal frequency can be passed while harmonics and the spurious signals are attenuated.

8-3.1.1. Shunt Capacitive Filters and General Capacitor Characteristics. The simplest low-pass EMI filter is a shunt capacitor connected from the interference-carrying conductor to ground. It serves to bypass high frequency (HF) energy, as indicated in the ideal representation of figure 8-1.

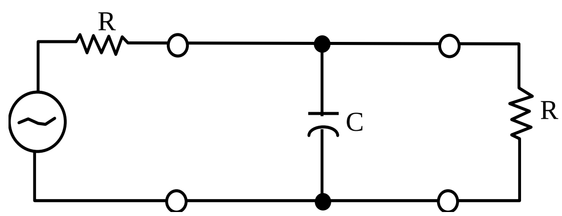


Figure 8-1. Capacitor Low-Pass Filter

Under these circumstances, the insertion loss of a shunt capacitor filter is defined by the relationship:

$$IL(dB) = 10 \log \ (1 + F^2) \qquad (8\text{-}2)$$

Where:

$F = \pi f R C$

f = frequency, in Hertz

R = driving or termination resistance, in ohms

C = filter capacitance, in farads

An actual capacitor incorporates both resistance and inductance. These effects are due to such factors as the foil inductance of the capacitor plates, lead inductance, foil resistance, and lead-to-foil contact resistance.

The variations in these inductive and resistive effects depend upon the type of capacitor. Metallized paper capacitors, while small in physical size, offer poor RF bypass capabilities because of high-resistance contact between the leads and the capacitor metal film. They are also a source of radio noise as the dielectric punctures and self-heals by burning away the metal film. This effect is indicated by the switch in the equivalent circuit shown in figure 8-2. The standard wound aluminum foil capacitor may be employed as a radio frequency bypass in the frequency range up to 10 MHz. Its useful frequency range of operation is a function of capacitance and lead length. Its equivalent circuit is the same as that of the metallized paper capacitor, but R_s and S of figure 8-2 are not in the circuit.

R_1 = lead-to-foil contact resistance

R = resistance of metallized foil

L = lead inductance

C_1 = capacitance

L = foil inductance

R_s = short circuit resistance

S = short circuit due to voltage puncture

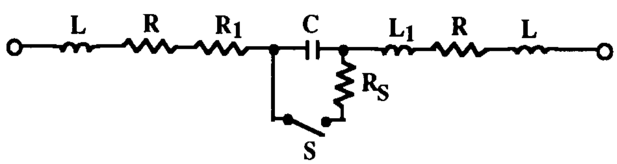


Figure 8-2. Metallized Capacitor Equivalent Circuit

As a result of these inductive effects, a capacitor will exhibit a capacitive reactance at low frequencies, and this situation will be maintained until its self-resonant frequency is reached. Above this frequency, the capacitor behaves like an inductive reactance. This effect is illustrated in figure 8-3. Also note the effect of changing capacitor lead length on this self-resonant frequency.

Mica and ceramic capacitors of small values are useful up to about 200 MHz. A capacitor of flat construction, if the capacitor plates are round as in a ceramic disc capacitor, will remain effective to higher frequencies than one of square or rectangular construction.

Other factors must be considered in selecting ceramic filter capacitors. A ceramic capacitor element is affected by operating voltage, current, frequency, age, and ambient temperature. The amount the capacity varies from its nominal value is determined by the composition of the ceramic dielectric. The dielectric composition can be adjusted to obtain a desired characteristic such as negative temperature coefficient, or minimum size. In obtaining one characteristic, other characteristics may become undesirable for certain applications. For example, when the dielectric composition is adjusted to produce minimum-size capacitors, the voltage characteristic may become negative to the extent that 50-percent capacity exists at full operating voltage, and full ambient temperature may cause an additional sizable reduction in capacity. Also, from the time of firing of the ceramic, the dielectric constant of the materials used may decrease; after 1000 hours, the capacitance may be as low as 75 percent of the original value. The designer should make ceramic capacitor selection based on required capacity under the most adverse operating conditions, and take into account aging effects.

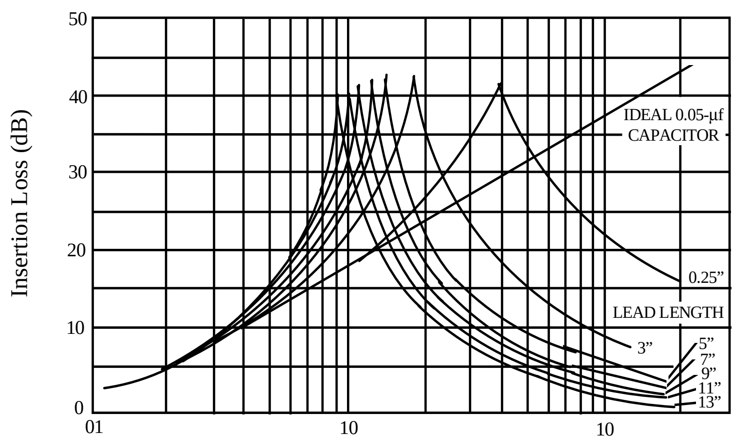


Figure 8-3. Insertion Loss of an 0.05-µf. Aluminum Foil Shunt Capacitor

Capacitors of short-lead construction, and feed-through capacitors, are three-terminal capacitors designed to reduce inherent end lead inductances. Figure 8-4 shows the construction of these three-terminal types. In each case, the inductance of the lead is not included in the shunt circuit. The wound foil short-lead capacitor is made with an extended foil-type construction so that each plate of the capacitor can be soldered to a washer-shaped terminal. One washer is, in turn, soldered to the center lead, while the other is soldered to the case that is the ground terminal.

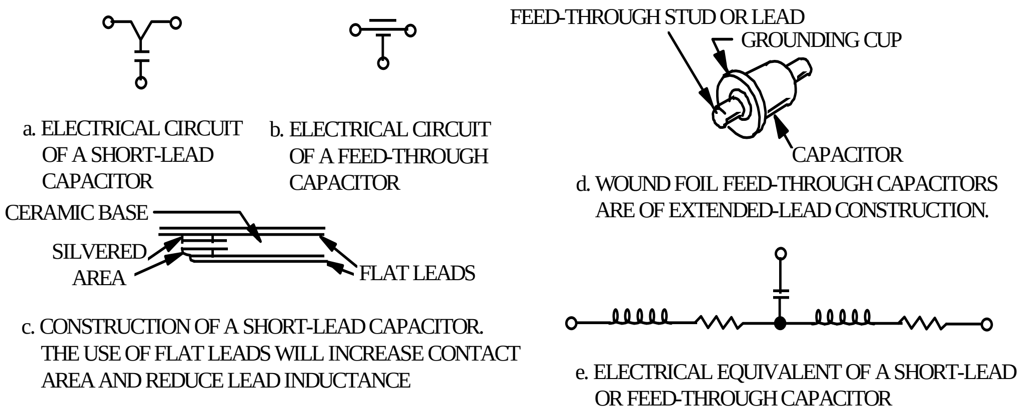


Figure 8-4. Three-Terminal Capacitor Construction

Theoretical insertion loss of three-terminal capacitors is the same as for an ideal two-terminal capacitor. However, the insertion loss of a practical three-terminal capacitor follows the ideal curve much more closely than does a two-terminal capacitor. The useful frequency range of a feed-through capacitor is improved further by its case construction, enabling a bulkhead or shield to isolate the input and output terminals from each other.

While the short-lead construction capacitor is ideally suited for EMI suppression in the frequency range of 1 to 1000 MHz, feed-through capacitors are available with a resonant frequency well above 1 GHz. The feed-through current rating is determined by the stud diameter. Figure 8-5 shows the insertion loss characteristics of typical feed-through capacitors.

Capacitor selection of shunt capacitive filters, or any other filter application, is determined in part by the voltage, temperature, and frequency range in which the filter must operate. For 28 V dc applications, capacitors rated at 100 working volts dc (WVDC) are adequate. Metallized mylar capacitors offer the most compact design and good reliability. Their dissipation factor is very low, and lead length can generally be kept short to improve HF performance.

Wet-type electrolytic capacitors are used for dc filtering and sometimes in EMI filters. They are single polarity devices, and their high dissipation factor or series resistance make them poor RF filters. An RF bypass capacitor should be placed across the output of dc supplies using electrolytics. The dissipation factors of electrolytic capacitors increase, and their capacitances decrease with age.

If a large value of capacitance is required in a small space, tantalum capacitors may be considered. Because tantalum capacitors are electrolytics, they are more sensitive to over-voltages, and are damaged by reverse polarity. The dissipation factor is considerably higher than for mylar or paper capacitors, and HF characteristics are poor. A large tantalum capacitor reaches its minimum impedance at 2 to 5 MHz or less, depending upon construction and capacitance value.

Capacitors for 120 V ac applications should be rated at 400 WVDC and be suitable for ac use. A unit of mylar and foil or of paper-mylar and foil is recommended. Dissipation factor is low and HF performance is good. For 240 V ac applications, an oil-impregnated paper and foil unit is recommended.

If good capacitor performance is to be expected above about 50 MHz, it is necessary to use designs incorporating feed-through techniques. As noted previously, lead inductance in a feed-through capacitor is not part of the shunt circuit, so that, compared with leads, it insertion loss is not degraded as rapidly with increase in frequency.

8-3.1.2. Series Inductive Filters and General Inductor Characteristics. Another simple form of low-pass filters is an inductor connected in series with the interference carrying conductor. It is ideally represented in figure 8-6. In practice, its insertion loss can be defined by the relationship:

$$IL(dB) = 10 \log \ (1 + F^2) \qquad (8\text{-}3)$$

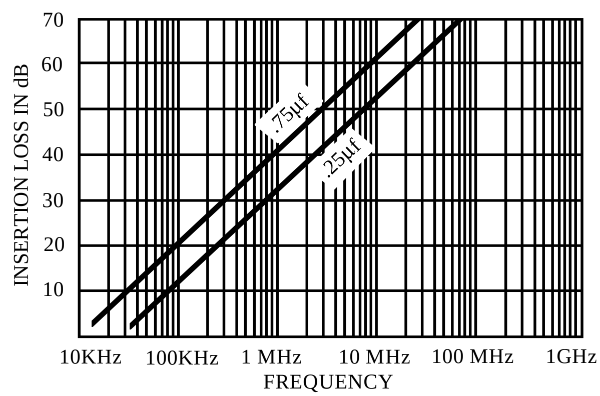


Figure 8-5. Typical Feedthrough Capacitor Insertion Loss

where:

$$F = \pi f \frac{L}{R}$$

F = frequency, in Hertz

R = driving or termination resistance, in ohms

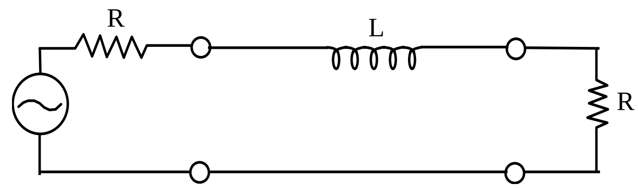


Figure 8-6. Inductor Low-Pass Filter

In practice, an inductor exhibits inductive reactance only until its self-resonant frequency is reached. Above self-resonance, it appears as a capacitive reactance, with the interwinding capacitance becoming dominant.

Filter inductors are usually toroidal, wound on cores of powdered iron, molybdenum permalloy, or ferrite material. The size of the core is determined by required inductance and current rating. The magnetic flux (number of turns multiplied by the peak current) should not drive the core to more than 50 percent of magnetic saturation.

The choice of core materials is determined by operating frequency and current rating. Powdered iron cores can be used for all dc applications and for most 60-Hz applications. For high-current 60-Hz devices, and for all 400-Hz applications, molybdenum permalloy cores should be used. Ferrite materials can be considered when the current that will flow through the inductor will not saturate the core.

Stray or distributed capacitance in a filter inductor has two detrimental effects: EMI may be coupled from input to output of the filter via the capacitance, and the capacitance may cause the filter to become self-resonant at one or more critical frequencies. Windings should be placed on the coil so that input and output turns are separated as much as possible to keep stray capacitance low. Distributed capacitance effects may be reduced by a careful arrangement of turns. In some cases, two or more coils wound on separate cores are connected in series to raise the self-resonant frequency.

Coil loss resistance is a measure of all power losses, hysteresis losses, and frequency-dependent absorption losses in the core. Loss resistance increases with frequency because of skin effect in the conductor, and due to changes in core loss with frequency. An increase in loss resistance represents an increase in attenuation in the filter passband. Losses in the core are not particularly detrimental, except that the insertion loss in the passband must be kept low.

8-3.1.3. Low-Pass “L” Section Filters. A primary disadvantage of single-element filters is that their out-of-band falloff rate is only 6 dB per frequency octave (20 dB per decade). By combining both a shunt capacitor and a series inductance single-element filter into an “L” configuration, a falloff rate of 12 dB per frequency octave can be obtained.

The two possible representations of low-pass “L” section filters are shown in figure 8-7. In 8-7a, the capacitor shunts the source impedance, while in 8-7b, the capacitor shunts the load impedance. The insertion loss for the “L” section filter is independent of the direction of inserting the “L” section into the line, if source and load impedances are equal. When source and load impedance are not equal, the greatest insertion loss will usually be achieved when the capacitor shunts the higher impedance.

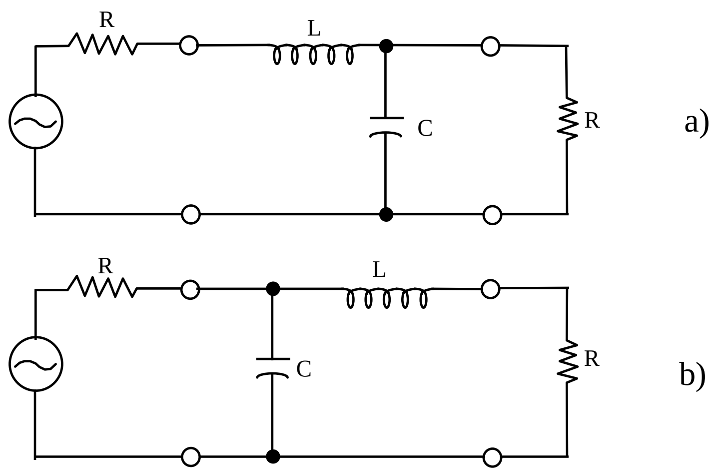


Figure 8-7. Low-Pass “L” Section Filter

The insertion loss of an “L” section lumped-constant network into 50-ohm resistance source and load impedances is:

$$IL(dB) = 10\log\left(1 + \frac{F^2D^2}{2} + F^4\right) \qquad (8\text{-}4)$$

where:

$$D = \frac{1-d}{\sqrt{d}}$$

$d = L/CR^2$ = damping ratio

$$\omega_\circ = \frac{\sqrt{2}(R)}{L} = \frac{\sqrt{2}}{RC}(if\ \ d = 1)$$

$$\omega_\circ\sqrt{2/LC}(if\ \ d \neq 1)$$

$F = \omega/\omega_\circ = f/f_\circ$ = normalized frequency

$\omega = 2\pi f$ = angular frequency (radians/second)

The “damping ratio,” d, relates the magnitudes of the filter elements to the magnitudes of the source and load impedances. It is defined so that setting d equal to one (ideal damping) results in the elimination of the squared frequency term from the insertion loss equation and produces an abrupt transition from the pass-band to the stop region. The equations for Butterworth filter designs are obtained when d is set equal to one.

Values of d less than one result in insertion loss curves identical to those obtained when d is greater than one. That is, for two element filters:

$IL(for\ d = n) = IL(for\ d = 1/n)$, for n = any real number (8-5)

The insertion loss of a two-element filter is not changed when it is “turned around” so that the source and load terminals are transposed, so long as the source and load impedances are equal.

The physical size of an “L”section filter depends upon insertion loss requirement, current rating, and voltage rating, with the first two usually predominant. The “L” section type of filter may give poor HF attenuation because of stray inter-turn capacitance. In some cases, the “L” type may resonate and oscillate when excited by transients.

8-3.1.4. π-Section Filters. The “π-“ section filter is the most common type of radio frequency interference suppression network. Figure 8-8 shows the circuit of the π-section filter. Advantages are ease of manufacture, high-insertion loss over a wide frequency range, and moderate space requirements. Although voltage rating must be considered, current rating and attenuation are the most important factors in determining the size of the filter.

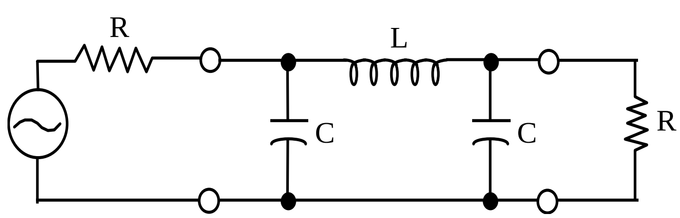


Figure 8-8. Low-Pass π Filter

The insertion loss of a lossless π-section network operating with equal source and load impedance is:

$$IL(db) = 10\ \log(1 + F^2D^2 - 2F^4D + F^6) \qquad (8\text{-}6)$$

where:

$$D = \frac{1-d}{\sqrt[3]{d}}$$

$d = L/CR^2$ = damping factor

$$\omega_\circ = \sqrt{2/LC} = 2R/L = 1/RC(if\ d = 1)$$

$$\omega_\circ = 3\sqrt{2/RLC^2}(if\ d \neq 1)$$

$$F = \omega/\omega_\circ$$

$\omega = 2\pi f$ = angular frequency

Unlike the “L” section filter case, overdamping or underdamping of π-section and (T-section) filters result in entirely different effects. This is discussed further in paragraph 8-3.1.6.

A typical attenuation curve of a π-section filter has a slope of approximately 18 dB per octave; the HF performance can be improved by internal shielding within the filter case. However, the “π” circuit is very susceptible to oscillatory ringing when excited by a transient.

The multiple π-section filter (cascaded π-sections) has characteristics identical to those of the multiple L-section filter. The attenuation curve of the theoretical multiple π-section filter rises at a rate of 20 dB more per decade of frequency than does a multiple “L” filter of the same number of sections. Although this may not be a significant factor when three or more sections are used, it does provide a capacitive input at both ends of the filter that is sometimes advantageous. An extensive use for this type of network is as a power-line filter in large installations, and for shielded enclosures where high attenuation is needed at very low frequencies.

8-3.1.5. <u>“T” Section Filters</u>. The “L” type low-pass filter can also be improved by the introduction of another series inductor. This addition forms a “T” section filter, which consists of two inductors in series with a shunt capacitor connected from the junction of the two inductors to ground (see figure 8-9).

Insertion loss is given by:

$$IL\ (dB) = 10\ \log(1 + F^2D^2 - 2F^4D + F^6) \qquad (8\text{-}7)$$

where:

$$D = \frac{1-d}{\sqrt[3]{d}}$$

$d = R^2 C/2L$ = damping factor

$$\omega_\circ \sqrt{2/LC} = 2R/L = 1/RC (\mathit{if}\ d = 1)$$

$$\omega_\circ = 3\sqrt{2/RLC^2} (\mathit{if}\ \ d \neq 1)$$

$F = \omega / \omega_\circ$

$\omega = 2\pi f$

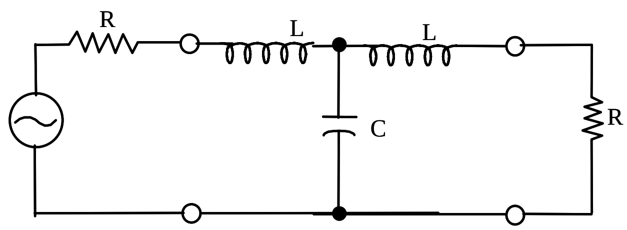


Figure 8-9. Low-Pass "T" Filter

The "T" type of filter is a very effective form of the lumped-constant type of filter for reducing transient interference. Its major disadvantage is the requirement for two inductors, which under some circumstances may present a size penalty. It provides the same out-of-band falloff rate as the π-section filter; that is, 18 dB/octave (60 dB decade) for a single section.

Figure 8-10 provides representative information on commercially available T-section, low-pass filters.

RATINGS

PART NO.	CURRENT AMPS	DCR AT 25°C OHMS	INSERTION LOSS CURVE	L_1 MAX INCHES	L_2 MAX INCHES
GF51F3A	0.10	3.20	A	.99	.94
GF51F3B	0.50	.66	B	.99	.94
GF51F3C	1	.36	C	.99	.94
GF51F3D	5	.022	D	.99	.94

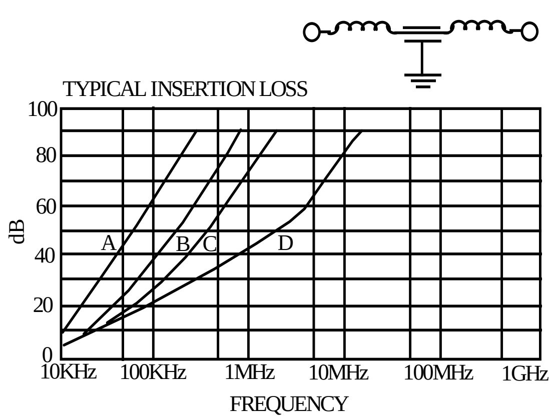


Figure 8-10. Representative Commercial Low-Pass T-Section Filter Characteristics

8-3.1.6. Insertion Loss Calculations for “π” and “T” Section Filters. The equations for the insertion loss of a T-circuit and a π-circuit as given by equations (8-6) and (8-7) are of the same form, but differ with respect to the definition of the equation parameters. The equation has three modes of response. When d equals one, the response is optimally damped and is the ideal (Butterworth) response curve. When d is greater than one, the response is in an overdamped mode. When d is less than one, the response is in an underdamped mode. In the underdamped case, the curve has a maximum in band loss of:

$$IL = 10\ \log(1 + 4D^3 / 27) \qquad (8\text{-}8)$$

at the frequency where $F = \sqrt{D/3}$. A minimum loss point will also occur at the frequency where $F = \sqrt{D}$.

8-3.2. LOSSY LINE FILTERS. While the input and output impedances of some filters can be expected to match their intended source and load impedances over a fairly broad frequency range, it is more often the case that such matches will not occur. For example, the input impedance of a powerline filter almost never achieves a match with the impedance of its associated powerline. As another example, a transmitter harmonic filter is generally designed to match the transmitter output stage over the fundamental frequency range, but not necessarily at its harmonic frequencies.

Because of such mismatch situations, there have been many cases when the insertion of a filter into a line carrying interference has actually resulted in more, rather than less, interference voltage appearing on the line beyond the point of its application. This deficiency in all filters composed of low-loss elements has let to the development of dissipative filters that take advantage of the loss-versus-frequency characteristics of magnetic materials such as ferrites.

Ferrites are inert ceramics containing granulated iron compounds. They are free of any organic substances, and are not degraded by most environments. Ferrites have both a large relative permeability (μ_r) and a large relative permittivity (ε_r). When $\mu_r = \varepsilon_r$ for a given ferrite, the wave impedance of the material is equal to the wave impedance of free space, and an impinging EM wave is absorbed without reflection. This occurs over a small relative bandwidth and is the reason for the loss-versus-frequency characteristics of ferrites.

One form of dissipative filter uses a short length of ferrite tube with conductive silver coating deposited on the inner and outer surfaces to form the conductors of a coaxial transmission line. The line becomes extremely lossy at radio frequencies; that is, it has high attenuation per unit length in the frequency range where either electric or magnetic losses, or both, become large. An example of the performance of a lossy line filter of this type is shown in figure 8-11.

Dissipative filters of this type are necessarily low-pass. One of the large uses of such filters is in general-purpose power-line filtering, in which the dissipative filter is combined with conventional low-loss elements to obtain the necessary low cutoff frequency.

Another method of achieving a dissipative filter is by use of lossy beads. Tubular ferrite toroids offer a simple, economical method for attenuating unwanted HF noise or oscillations. One bead slipped over a wire produces a single-turn RF choke that possesses low impedance at low frequencies and moderately high impedance over a wide HF band.

The presence of a ferrite bead on the wire causes a local increase of series impedance (largely resistive) presented to currents in the wire. Figure 8-12 illustrates the effects of one ferrite bead on a length of wire. Adding more or longer beads provides additional units of series inductance and resistance in direct proportion. Extra turns of wire can be passed through the bead, increasing both resistance and inductance in proportion to the square of the number of turns. Because of distributed winding capacitance, this technique is most effective at the lower frequencies. Because of the high resistivity of ferrite beads, they may be considered insulators for most applications.

High-amplitude signals below 50 MHz may cause some reduction in the suppression effect due to ferrite saturation. However, as long as only one turn links the core, fairly high currents can be tolerated using representative materials before saturation is approached. At saturation, inductance and resistance will be low, but will return to normal values upon removal of the high field.

Still another form of ferrite filter that extends the ferrite bead concept is the filtering connector. Lossy filters are built directly into a male connector assembly, and offer low-pass filter performance as shown in figure 8-13.

Lossy line suppressant tubing also provides an efficient means for suppressing undesired EMI and other spurious signals. The tubing can be slipped over standard wire and cable and suppresses both radiated and conducted energy. It can be used in environments from -55 degrees to +250 degrees C without electrical or mechanical degradation.

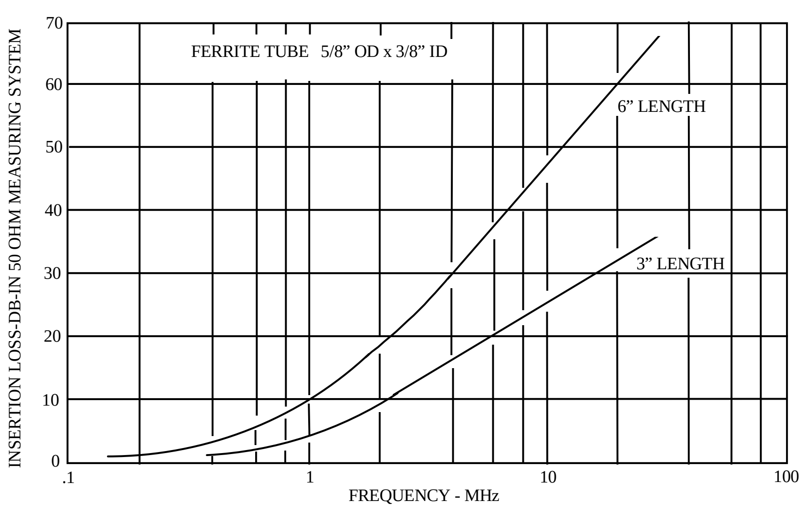
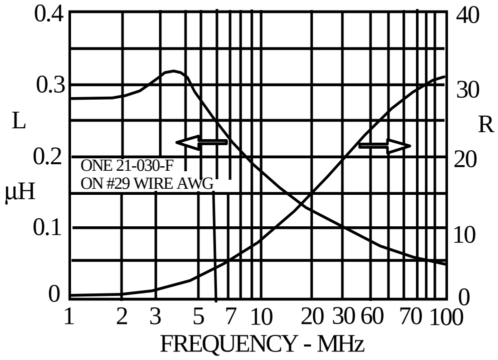


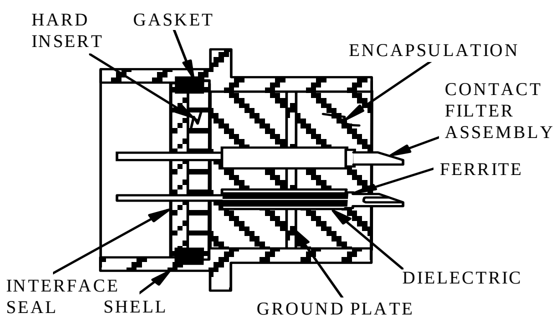
Figure 8-11. Insertion Loss of a Ferrite Tube Low-Pass EMI Filter

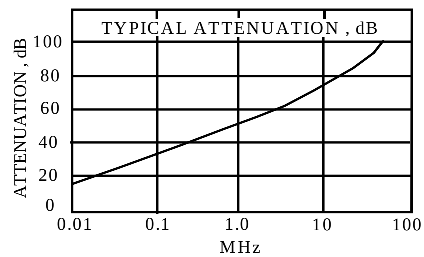


Typical values of equivalant series resistance and inductance attributed to ferrite beads.

TYPICAL PROPERTIES	
Flux Density (B) at 5 0_e	2400 G
Coercive force (Hc)	0.56 0_e
Hysteresis Factor (h/μ^2)	$22x10^{-6}$
Initial Permeability (μ0)	450
Permeability (μ) at 250G	900
Resistivity - Ohm - cm	≥ 10^7
Curie Temperature	≥ 156^oC

Figure 8-12. Filter Characteristics of Ferrite Beads





Minimum Attenuation from –55° to +125° and
100 MHz to 10 GHz ....80 dB
D.C. Working Voltage (includes summation of the D.C. and
low level A.C. superimposed peak voltage) ....50 VAC
Dielectric Strength (for 5 sec, with changing current of 50
milliamperes maximum) ....100 VDC
Feed Through Current (Nominal) D.C. and/or
audio frequency RmS ....7.5 Amperes
Insulation Resistance ....250 Megohms
R.F. Current ....0.25 Amperes
Operating Temperature Range ....-55 C to + 125 C
Capacitance (μf) ....1μf Nominal

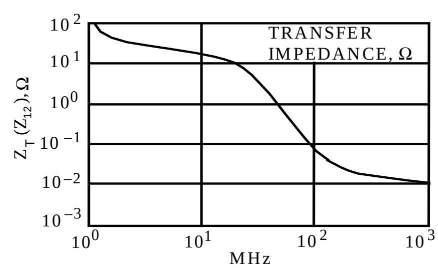


Figure 8-13. Typical Characteristics of Lossy Connector

Improvement of HF rejection characteristics of a conventional low-pass filter may be obtained by employing a conventional reactive filter in cascade with a lossy line section. This arrangement can provide an overall characteristic having both a rapid cutoff slope and a high-stop band attenuation. An example of the improvement in stop-band attenuation that can be gained by preceding a reactive filter with a lossy line section is illustrated in figure 8-14.

Figure 8-14A shows the performance of a reactive low-pass filter constructed with lumped constant elements. The rapid cutoff at 400 MHz is followed by a high-attenuation region between 400 MHz and 3 GHz, but at frequencies above 3 GHz the attenuation is greatly reduced. If the same low-pass filter is preceded by a section of coaxial line whose dielectric space is filled with a 6:1 radio of iron-to-epoxy dielectric material, the attenuation characteristic is altered to that shown in figure 8-14B. The addition of the lossy section has increased the pass band attenuation only slightly, but the stop band attenuation has been increased to greater than 60 dB.

When a lossy line section is used in cascade with a conventional low-pass filter, the pass band insertion loss can be minimized by the proper choice of the dielectric material. However, there is always some pass band loss introduced by the lossy dielectric. Such pass band losses can be reduced by designing the reactive filter to have as wide a region as possible between the low-pass cutoff frequency and the first spurious pass band, so that a minimum of lossy material is needed to provide the required stop band attenuation.

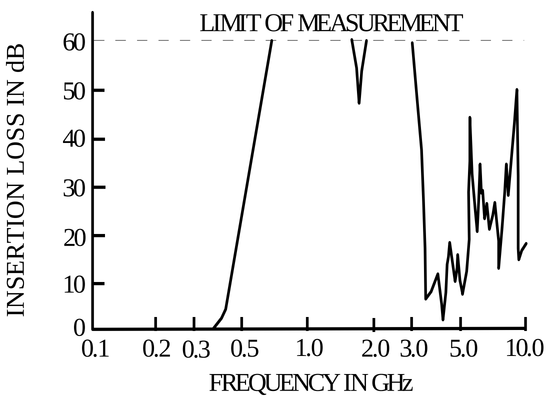


Figure 8-14A. Typical Low-Pass Filter Loss Characteristics, Low-Pass Filter Only

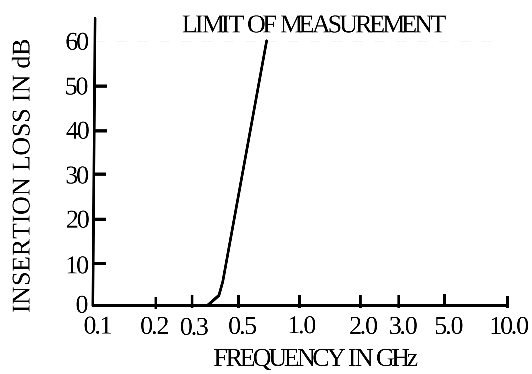


Figure 8-14B. Typical Low-Pass Filter Loss Characteristics, Plus Lossy Filter Section

8-4. TRANSIENT SUPPRESSION. The making and breaking of current, either mechanically or electronically, can introduce transient radiated and conducted effects. The transient may be generated within the equipment as a result of a switching function, or external to the unit, such as from other equipment tied to the same power line, or from the weapon system being tested.

If the source of the transient is external to the ordnance and cannot be reduced at its point of origin, then the shielding, bonding, grounding, and filtering techniques already discussed in this design guide must be applied by the designer to prevent the transient energy from affecting ordnance performance. If the source of the transient is within the equipment, the designer should be aware of particular switch transient suppression techniques that can be applied.

8-4.1. INDUCTIVE LOADS. When an inductive load is opened by a switch contact, such as when a relay coil is opened, a reverse voltage is produced by the collapsing magnetic field. This reverse voltage increases until an arc occurs across the contact. The arc produces a wide frequency band of interference which is conducted and radiated away from the switch. Shielding and filtering will be necessary to reduce noise conduction and re-radiation from the switch contacts and wiring.

Figure 8-15 shows a number of suppression techniques that can be used to minimize transients in circuits that switch inductive loads. These techniques are described in the following paragraphs:

8-4.1.1. <u>Resistance Damping</u>. Use of a non-inductive resistor across the load is the least expensive approach but it has drawbacks because the resistor will continuously dissipate power, and because the suppressor will increase relay dropout time. Maximum protection is obtained when the value of the suppression resistor, R, is equal to the coil resistance, R_L, but this will generally result in severe steady-state power consumption requirements. Practically, R is kept as low as possible consistent with power capabilities and dropout time. A typical resistance value chosen is 10 times R_L, but this size may cause the dropout time of the relay to increase by as much as a factor of 10.

8-4.1.2. <u>Capacitance Suppression</u>. A very popular relay suppression circuit consists of a series resistor capacitor combination placed across the coil. Care must be taken in this arrangement to avoid circuit resonance effects that will cause contact chatter, and to avoid overdamping so that dropout time does not become excessive. A typical arrangement is for R to be ¼ to ½ of R_L. The value of the capacitance, C, can then be obtained from the relationship:

$$C = \frac{L}{RR_L} \qquad (8\text{-}9)$$

where L is the inductance of the load.

If the computed capacitance is very large, use the closest practical smaller capacitor. Capacitance between .01 to 1 uf is usually sufficient to minimize most transient effects. The RC suppressor can create EMI problems when the relay is actuated, since there is an increase in current flow until the capacitor is charged. It is often preferable to place the suppressor across the relay contacts rather than across the coil to reduce this problem.

8-4.1.3. <u>Single Diode Suppression</u>. A series diode-resistor combination across the coil eliminates the power dissipation problem of the resistor suppressor. When the relay is energized from a dc source, the diode is back-biased so that the suppression circuit has no effect on relay on operation. When current to the relay coil is opened, a current flows through the suppression circuit; this current is generally a function of the value of the suppression resistor R (since the value of R is usually significantly larger than the forward resistance of the diode). The resistor value should be kept as small as possible consistent with relay dropout requirements.

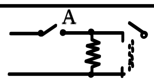
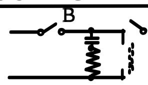
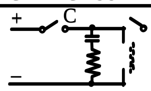
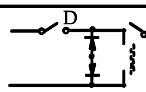
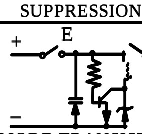
TYPES OF INDUCTIVE SUPPRESSION	VOLTAGE INPUT	RELAY CONTACTS		REMARKS
		CLOSING	DROPOUT	
 RESISTANCE DAMPING	AC or DC	NO EFFECT	FUNCTION OF RESISTANCE	Increase in power consumption. Resistance should be as low as practical. Observe power rating, E^2/R, and heat dissipation.
 CAPACITANCE SUPPRESSION	AC or DC	SLIGHT EFFECT	SLIGHT EFFECT	Need series resistance of a few ohms. Capacitance value around .01 to 1 μf. Capacitance rated 10 times the input voltage.
 SINGLE DIODE SUPPRESSION	DC ONLY	NO EFFECT	SLIGHT EFFECT	Polarity is critical; diode is connected in backward or nonconductive direction. PIV should be higher than any transient voltage plus safety factor. Series resistance of a few ohms might be needed to increase inductive life.
 BACK-TO-BACK DIODE SUPPRESSION	AC	NO EFFECT	NO EFFECT	Avalanche voltage should be above input voltage. Power dissipation should be sufficient for transient current. Cost of device becoming a significant factor.
 DIODE-TRANSISTOR SUPPRESSION	DC	NO EFFECT	NO EFFECT	Most effective in transient suppression. Voltage transient and dropout time negligible. Most expensive technique.

Figure 8-15. Comparison of Various Suppression Devices across an Inductor

Silicon diodes are used for this application because of their low cost, adequate current ratings, and ability to meet peak inverse voltage (PIV) requirements. Zener diodes also have been employed for this purpose.

8-4.1.4. Dual Diode Back-To-Back Suppression. An extension to the single diode suppression approach is the use of back-to-back diodes across the relay coil. This arrangement is insensitive to the polarity of the relay control signal and can therefore be employed with ac relays. The component that conducts when power to the relay coil is removed may be a Zener diode, to clamp the transient at a specific level. Use of a Zener diode can help the dropout time problem as well.

8-4.1.5. Diode-Transistor Suppression. A very effective transient suppression approach is to use two diodes, a resistor, and a transistor in an arrangement as shown in figure 8-15E. When the switch is opened, the transistor is cut off, forcing the transient energy to dissipate through the Zener diode Z and the conventional diode D. While this suppression circuit results in good transient control, it is a relatively expensive alternative, and should only be considered when the previously discussed approaches are not suitable.

8-5. FILTER INSTALLATION AND MOUNTING.

When filters are used, it is absolutely necessary to follow certain installation guidelines if good results are to be obtained. The RF impedance between case and ground must be kept as low as possible. Otherwise, the filter insertion loss will be seriously degraded at the higher frequencies. Effective separation of input and output wiring is mandatory because the radiation from wires carrying interference signals can couple directly in output wiring, thus circumventing and nullifying the effects of shielding and filtering. If complete isolation is affected between input and output, filter insertion loss will approach the design figure.

Both filter mounting guidelines given above can be readily satisfied by the use of bulkhead mounted feedthrough filters such as shown in figure 8-16. A wide variety of these filters is available as off-the-shelf items from a number of manufacturers that can satisfy most filtering requirements.

The preferred method of mounting feedthrough filters in equipment that must be used in a high-level EME (such as an aircraft carrier flight deck) are to mount the filters in a metal enclosure behind the front panel (known as a "doghouse").

This enclosure can be constructed with an access panel on one of the sides. However, the panel should be attached to the doghouse using an *RF* gasket following the manufacturer's recommendations for obtaining an effective seal.

Another effective filtering method uses a filtered pin connector. In this case, the EMI filter is an integral part of the connector. The filtered pin connector will generally require less equipment space for mounting than will a connector/filter/doghouse assembly.

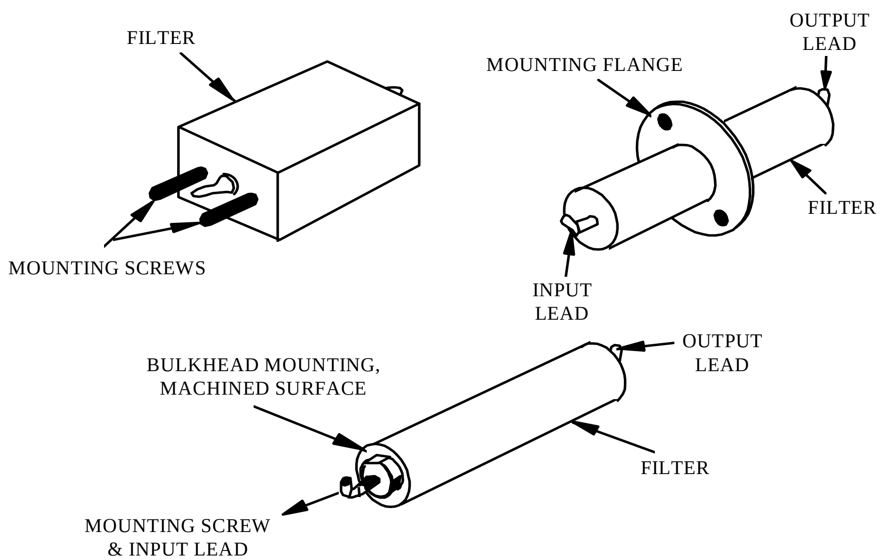


Figure 8-16. Typical Feedthrough Filters for Bulkhead Mounting

8-6. SPECIFYING FILTERS.

In designing or selecting a filter for a particular application, many parameters must be taken into account if the filter is to be effective. Its insertion loss versus frequency curve is obviously the primary characteristic that determines the suitability of the filter for a particular EMI application. However, other electrical and mechanical requirements must also be designated. They are discussed in the following paragraphs:

8-6.1. IMPEDANCE MATCHING. The elements of the filter must be chosen so that the impedance network matches the line into which it is inserted. This is especially true of the transmission lines, so that the filter does not impair the normal function of the equipment at both ends of this line. When a filter must be installed in a circuit where its source impedance and/or load impedance is either not known or may vary over a relatively wide range, it may be desirable to terminate the filter into fixed impedances to stabilize its performance.

8-6.2. VOLTAGE RATING. Consider the voltage rating on the filter, particularly if used on power lines. Under some conditions, the voltage may deviate by a large amount from its normal value. In addition, short duration pulses whose amplitudes are well above rated line voltage may be on the power circuits, both AC and DC. The filter voltage rating must be sufficient to provide reliable operation under the extreme considerations expected.

8-6.3. VOLTAGE DROP. Determine the maximum allowable voltage drop through the filter and design accordingly.

8-6.4. CURRENT RATING. Current rating should be for maximum allowable continuous operation of the filter. Calculate the current rating for filter elements, such as capacitors, inductors, and resistors. Whenever possible, the current rating of filters should be consistent with the current rating of the wire, circuit breakers, or fuse with which the filter will be used. A filter with a higher current rating than the circuit in which it is installed will often add a weight and space penalty. A filter with a lower rating will have poor reliability and maybe a safety hazard. The safety used in rating filters should also be consistent with those used for other circuit components.

8-6.5. FREQUENCY. Consider both the operating frequencies of the circuit and the frequencies to be attenuated. In general, the cost of a filter rises rapidly as the required rate of skirt falloff goes up, so care should be taken in identifying insertion loss versus frequency needs. Also, select the appropriate type of filter - a band-reject filter might be better to reduce the level of a single close-in narrowband source, rather than to use a sharp falloff low-pass filter.

8-6.6. INSULATION RESISTANCE. The insulation resistance of the filter may vary during the life of the filter. Determine the maximum allowable variation of this resistance for proper filter operation, and design accordingly.

8-6.7. SIZE AND WEIGHT. Size and weight may be important in some GSE filter applications. When space is at a premium, adding or subtracting various filter elements may be traded against reduced size and weight of the filter. Filter manufacturers are fairly flexible in being able to provide a wide choice in shape of the filter unit, its method of mounting, and the methods of making connections.

8-6.8. TEMPERATURE. The filter must be able to withstand the environmental operating ranges of the equipment in which it will be used. In most cases, the GSE requirement will probably necessitate a -65 to +85 degree C temperature range.

8-6.9. RELIABILITY. Filter component reliability must be commensurate with the equipment reliability requirement, and should be high, relative to other equipment components. This is primarily dictated by the fact that faults in EMI filters may be somewhat more difficult to locate than faults in other components.

As an aid to filter procurement, table 8-2 is provided. This type of form is similar to that contained in the appendix of MIL-F-18327C, but is applicable to both MIL-F-18327C and MIL-F-15733E requirements. It can be useful to the designer in establishing filter requirements, whether they are to be met in-house or by an outside source. The designer who must specify EMI filter requirements should be familiar with MIL-F-18327C and MIL-F-15733E (as well as MIL-STD-2020, not only as far as table 8-2 is concerned), but to obtain a clear understanding of the requirements that must be imposed on filters.

8-6.10. LEAKAGE CURRENT (POWER LINE FILTERS). The capacitance between individual inputs to a power line filter and the filter case should be minimized to control leakage currents on the ground return. Low leakage will prevent shock hazards to personnel if the filter case should become ungrounded, and will also reduce common-mode interference (hum) as well as prevent inadvertent tripping of ground fault indicating devices. Maximum filter leakage current in a single equipment should not exceed 30mA per phase (5mA per phase in test equipment) in order to comply with existing specifications. This translates to $0.6\mu f$ and $0.1\mu f$ ($0.1\mu f$ and $0.016\mu f$ for test equipment) of input capacitance per phase at 60 Hz, and 400 Hz, respectively. Where filtering requirements cannot be met without exceeding the input capacitance limits, an isolation transformer should be used to isolate the equipment from the power source. Other means such as grounding the powerline neutral at the filter case should be explored.

8-7. TERMINAL PROTECTION DEVICES.

A number of devices have been developed which can aid in the protection of electrical penetrations into the shielded compartments of an ordnance system. Electrical transients on signal, power, control, and other interfacing cables can upset, damage, or even activate electrical circuits which could initiate HERO events. This section describes a number of the terminal protection devices (TPD's) that can be used to mitigate overvoltage or induced transients at entry points into the ordnance. The most often used TPDs are metal oxide varistors (MOV's), avalanche (Zener) diodes, gas discharge tubes, and thyristors.

Table 8-2. Filter Information Sheet

Duration of test voltage (if other than 60 seconds) _____ seconds.
Points of application of test voltage _____.
At reduced barometric pressure: Method 105, MIL-STD-2020.
Test condition B _____. Other (specify) _____.

Insulation resistance: Method 302, MIL-STD-2020.
Test condition A _____ Other (specify) _____.

Insulation resistance: Method 302, MIL-STD-2020.
Test condition A _____ Other (specify) _____.
Points of measurement: Between terminal and mounting _____, between terminal and case _____, Other (specify) _____.

Overload: Paragraph 4.6.10, MIL-F-15733E _____.

Stability at temperature extremes: Paragraph 4.7.6, MIL-F-18327C.

Life: Method 108, MIL-STD-2020.
Test condition B _____, Test condition D _____, Test condition F _____.
Duration of test if other than test condition B,D, & F _____ hrs.

Operating conditions: Test potential _____, Duty cycle _____, Load _____,
Test Temp. _____, Tolerance _____.

Temperature rise: Paragraph 4.7.9, MIL-F-18327C _____. Paragraph 4.5.4, MIL-F-15733E _____.

Vibration: Low frequency _____, Method 201, MIL-STD-2020. Load condition _____.

Shock: Method 205, MIL-STD-2020, Paragraph 4.7.12, MIL-F-18327C.

Temperature cycling: Method 104, Condition A, MIL-STD-2020, and Paragraph 4.5.15.2, MIL-F-15733E _____.

Immersion: Method 104, MIL-STD-2020, Test condition A _____, B _____, C _____.

Moisture resistance: Method 106, MIL-STD-2020, Paragraph 4.7.14, MIL-F18327C _____
.

Flammability (grades 5 and 7): Method 111, MIL-STD-2020.

Salt Spray: Method 101 of MIL-STD-2020.

Table 8-2. Filter Information Sheet (Continued)

<table>
<tr><td colspan="3">Electrical characteristics

Impedance: Input _____ ohms. Output _____ ohms. Transfer _____ + _____

ohms (if applicable). Load _____ ohms.

Resonating Capacity _____ Pf + _____Pf (if applicable).

Capacitance to ground: Method 305, MIL-STD-2020 and 4.6.3, MIL-F-15733E:
_____ Pf at _____ kHz, _____ pf at _____ at _____ kHz,

Single input voltage _____ volts.

Current rating: AC, _____ amps DC, _____ amps

Voltage rating: AC, _____ volts DC, _____ volts.

Duty Cycle (if intermittent): Time on _____ Time off _____

Maximum allowable voltage drop: _____ volts.

Reference frequency _____ (Hz) _____ (kHz) _____ (MHz) _____.

Insertion loss (at reference frequency) _____ dB (min).</td></tr>
<tr><td>Discrimination
Frequency Range

_____ (Hz) _____ (kHz) _____ (MHz)</td><td>Min
(dB)</td><td>Max
(dB)</td></tr>
<tr><td></td><td></td><td></td></tr>
</table>

8-7.1. MOV'S. Metal oxide varistors are made from sintered metal oxides, primarily zinc, and act as non-linear resistances. When the MOV is exposed to a voltage greater than the rating of the device, the varistor's resistance changes to a low value allowing the MOV to absorb the transient energy. Once the voltage is removed, the MOV will return to its state of high resistance.

MOV energy dissipation capability is determined by its physical size and can be very robust in power ratings. A disk on the order of 20 mm can withstand a one-time pulse of up to 6 kA. Another advantage of using a MOV is that because it is a clamping device, it can be used across an ac line.

The disadvantage of the MOV is that its capability is size-dependent and devices less than about 14 mm are not effective. Because it is a clamping device, the MOV must absorb a tremendous amount of energy which breaks down the internal structure of the device and causes the MOV to eventually fatigue. As it fatigues, its performance degrades until the device eventually fails. This, coupled with the slow response time of the device, prevents the MOV from being able to protect sensitive electronics.

8-7.2. AVALANCHE DIODES. Avalanche diodes are basically back-to-back Zener diodes that limit the voltage across a component or line, by clamping the incoming transient voltage after the stand-off voltage is exceeded. Much like a MOV, the avalanche diode is limited by the energy it can absorb while clamping the transient. But unlike a MOV, the avalanche diode does not have the same size (mass) and is typically restricted to low-voltage transients in order to limit the power that the silicon device must dissipate.

The advantages of using an avalanche diode are its speed and repeatability. Since it is made of silicon, the diode is much faster than the MOV or a gas discharge tube, and its capability will not degrade with time, unless it is stressed beyond its power rating. The design restriction for avalanche diodes is that they are both voltage- and current-dependent. Therefore, as the voltage requirement increases, the current capability will decrease.

8-7.3. GAS DISCHARGE TUBES. Gas discharge tubes are glass or ceramic tubes filled with an inert gas and terminated on each end with an electrode. The gas discharge tube is connected in parallel with the circuit being protected and has a very high off-state impedance. Once an incoming transient exceeds the dc breakdown voltage of the tube, the tube fires, causing an arc in the tube. This arc ionizes the inert gas which provides a low impedance path for the transient. When the transient voltage drops below the dc hold voltage, the gas discharge tube returns to it high off-state impedance.

An advantage of the gas discharge tube is that it can provide protection against a high surge current. For transients on the order of 10 microseconds, the gas discharge tube can withstand up to 20 kA. For longer, but lower amperage pulses, the gas discharge tube can handle many hundreds of transients.

A disadvantage of the gas discharge tube is that the performance does degrade with each transient. This occurs because the electrodes become contaminated with carbon, which increases the firing voltage of the device. The gas tube also alows a large overvoltage on the line while the inert gas is ionizing. Overvoltage can be on the order of four times the rating of the tube.

8-7.4. THYRISTORS. The thyristor is a solid-state device that operates very much like a gas discharge tube in that it acts like a crowbar. In the off state, the thyristor appears as a very high impedance. Upon application of a transient voltage exceeding the breakover voltage of the device, the thyristor will short to a very low resistance. The device will stay in this state until the current is either interrupted or drops to a level below the holding current of the device. At that time, the thyristor will reset itself to the high-impedance state.

Advantages of the solid-state thyristor are the speed of its activation and its reliability. There is very little overshoot on the transient, since the thyristor fires very rapidly. Also, the thyristor does not fatigue and can take many transient hits without degradation.

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CHAPTER 9

MANAGEMENT

9-1. HERO MANAGEMENT PROGRAM.

The success in controlling Hazards of Electromagnetic Radiation to Ordnance (HERO) during weapon system design and development is often dependent on the extent to which an electromagnetic compatibility (EMC) Management Program is established and implemented. Military Handbook 237, "Electromagnetic Compatibility Management Guide For Platforms, Systems and Equipment," is approved for use by all departments and agencies of the Department of Defense, and is written to provide managers with guidance for establishing an effective EMC Program throughout the life cycle of platforms, systems, subsystems, and equipment. Experience has shown that the avoidance of HERO problems, and EMC problems in general, can best be achieved by adhering to proper EMC design, development, test, production and installation requirements, practices, and procedures. To accomplish this, an effective program of EMC management, assessment, engineering and configuration control is required and must be integrated into the overall design and engineering effort from early in the conceptual phase and throughout the life cycle.

9-2. OBJECTIVE.

The objective of the HERO program is to certify that all fleet systems containing electrically initiated devices (EID's) are safe and reliable from stockpile-to-safe separation when exposed to the broad spectrum of electromagnetic environments (EME's) that will be encountered during their life cycle.

9-3. APPROACH.

In recognition of the fact that the best solution to HERO is a well-designed system, the HERO program has been structured in a coordinated manner to monitor system development from conception to retirement. The five basic phases in Naval system development as described in MIL-HDBK-237 are:

Phase 0 - Exploration and Definition (CED)
Phase I - Demonstration and Validation (D&V)
Phase II - Engineering and Manufacturing Development (EMD)
Phase III - Production and Development (P&D)
Phase IV - Operations and Support (OPS)

During Phase 0, the requirement to meet MIL-STD-464 is incorporated in the purchase description, system specification, or the EMC specification. This defines the following criteria for the system developer.

a. The EME that the system will be subjected to during its life cycle.

b. The pass/fail criteria for the system's EID's when subjected to a government-controlled predeployment test and evaluation. Also during this Phase, all prospective system contractors will be provided with a copy of this design guide which has been written to amplify and augment MIL-STD-464.

In Phase I, the system developer, now aware of HERO, should be prepared, with the aid of this design guide, to design in protection for the system's EID's. Available during this Phase, as well as throughout the system's life cycle, are the services of a HERO consultant. The consultant's task is to aid the system development in implementing the design principles and practices of this design guide. The HERO consultant should also serve as a member of the Electromagnetic Compatibility Advisory Board (EMCAB) for the system. During Phase I, test-article hardware requirements for quantities and configurations required to conduct the evaluation test will be presented to the developer.

Phase II is the phase in which the paper designs are first applied to hardware. At this time, the developer can make use of MIL-STD-1377. This document provides the system developer with methods of determining the effectiveness of the shielding and filtering that has been designed into the system to preclude HERO. During Phase II, a schedule for delivery of the test hardware and test dates will be established.

During Phase III, the system will be subjected to a government-controlled HERO evaluation test. The test will be performed on production hardware and will be conducted at a government EMC facility. The government EMC facility chosen for the evaluation will permit convenient and adequate simulation of operating shipboard EME's and will include a ground plane with appropriate radiation sources. HERO test facilities for HERO evaluation testing presently include three ground planes, shielded laboratory areas, mode-stirred chamber, and the equipment necessary for simulating the EME required to accomplish the tests. In cases where the ground plane simulation is not adequate; e.g., shipboard missile systems, the test will be conducted on the actual system platform on ship or at a land-based test facility.

The final Phase in system development is the Operations and Support (OPS) Phase. This is the phase in which the system is introduced to service as an operational entity capable of performing its intended mission. During the OPS Phase, system ECP's will be monitored and analyzed to determine if conditions exist that might require positive corrective action to maintain the HERO safety and reliability margins established during development. Follow-on developments and modifications of the system will be analyzed to determine if further testing is required to certify the modifications.

9-4. SUMMARY.

The HERO Management Program provides a well-founded methodical approach to systems EM safety from inception to retirement. This time-proven approach provides the Fleet with HERO-safe systems and avoids the time and money consuming process of retrofit.

CHAPTER 10

TESTING

10-1. GENERAL.

NAVSEA Instruction 8020.7 (series) requires that weapon systems and devices containing electrically-initiated devices (EID's) be analyzed and tested, if necessary, in order to obtain positive certification that they can be handled with impunity in the maximum predicted electromagnetic environment (EME) before they are issued to the Fleet. For most systems, this certification requires HERO evaluation tests.

10-2. PURPOSE.

The purpose of this chapter is to describe the nature and extent of the tests required for Navy certification, and to introduce tests that may be conducted by the developer to assist him in implementing HERO design requirements.

10-3. NAVY HERO CERTIFICATION.

The nature of weapon systems makes HERO testing mandatory for an operational system with the entire weapon system exposed to the EME (chapter 2). Navy HERO tests on both prototype and production models are normally conducted at a ground plane facility. In some instances, weapon launcher size or unique ship interfaces dictate that the test be performed aboard ship.

10-4. GROUND PLANE AND LABORATORY TEST FACILITIES.

In order to conduct the HERO test, ground plane facilities that permit convenient and adequate simulation of shipboard operational environments are required. These facilities include a ground plane of suitable size and location, with appropriate radiation sources. HERO test facilities for Navy certification tests presently include three ground planes, shielded laboratory areas, mode-stirred chamber, and the equipment necessary for simulating the EME required for Navy HERO tests.

The ground planes measure 100 x 240 feet and are constructed of welded steel plates. Figure 10-1 depicts a ground plane, its array of radiation sources, and a weapon system being tested.

In support of the ground plane test, tests may also conducted in the shielded areas of the laboratory and in the mode-stirred chamber in support of the ground plane tests. These laboratory tests allow component and subsystem tests, in addition to complete frequency coverage not possible on the ground plane.

10-5. PREPARATION OF THE WEAPON.

To measure the amount of electromagnetic energy induced from the environment in the weapon system=s EID’s, all explosives are removed from the weapon and radio-frequency (RF) sensing devices are placed near the intact bridgewires of the EID’s. These sensors ensure accurate measurement of absolute RF current values induced at each EID location, yielding a quantitative measure of weapon susceptibility to electromagnetic energy.

10-6. ENVIRONMENT FOR TEST.

The EME levels to which the weapon will be subjected are established prior to the test. Typical communication and radar equipment is used to develop these EME levels on the ground plane. Under all test conditions, the EME level is equivalent to the shipboard levels, or a known relationship exists that permits extrapolation of the test measurements to the shipboard EME.

10-7. TEST CONDITIONS AND PROCEDURES.

The test conditions and procedures used to evaluate a weapon system are designed to simulate the physical and electrical environment that will be encountered in shipboard operational situations. The frequency and radiated field parameters used in the tests are based on measurements taken aboard ship while the ship's communications and radar systems are operating.

The design of each weapon system includes specifications for the loading and handling procedures used for that particular weapon system in all operations aboard ship. These procedures use carts and cranes, the loading and unloading operations, the handling and connecting of cables, test and monitor functions on the aircraft and auxiliary equipment, safe-and-arm functions, and any other weapon launching operations. All procedures associated with the system are evaluated in the HERO certification of that system.

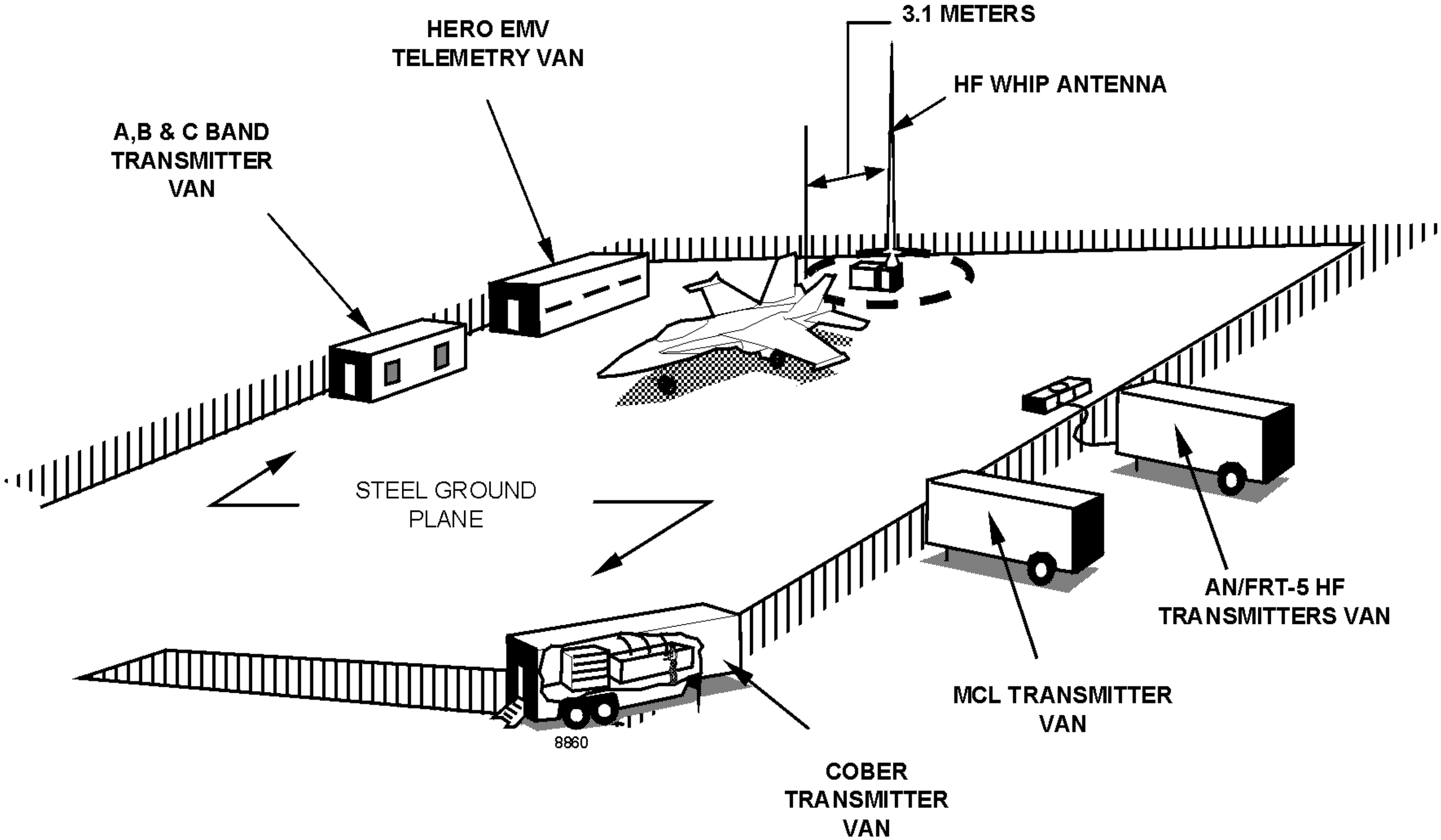

Figure 10-1. Ground Plane Test Facility, Naval Surface Warfare Center, Dahlgren Division

Some of the variables affecting the HERO characteristics of a weapon system, in addition to the handling and loading procedures previously mentioned, are frequency, field intensity, radiated power, weapon/aircraft orientation, and distance from the radiation source. Since there are many possible combinations of these variables, the tests are designed to examine specific conditions most relevant to the hazard while exercising control over the other less relevant factors, if possible. Examples of such factors are the proximity of personnel to the weapon, adjacent structures, variations in grounding or tiedowns, and improper application of test and checkout equipment.

The test results are used to determine the ordnance level of susceptibility in the expected shipboard EME. Additional tests are sometimes performed to further characterize the system=s degree of susceptibility. These situations include unconventional handling procedures or environments of higher electromagnetic energy levels. Observation of weapon system susceptibility under such conditions leads to the development of procedures that assure the safety and reliability of the weapon system throughout the operational life cycle.

10-8. PROTOTYPE VERSUS PRODUCTION WEAPONS TESTS.

The determination of an ordnance system's HERO susceptibility must not be considered an "after-the-fact" responsibility. A continuous assessment of HERO susceptibility throughout the design-prototype-production phases of development must be implemented. See chapter 9.

When the prototype systems have successfully passed the HERO test, any change, however insignificant it might seem, must be recognized as a potential problem area. If it is impossible (or undesirable in a performance sense) to maintain continuity from prototype to production phases, the design modification must ensure that the ordnance remains as safe and reliable as the tested prototype from a HERO standpoint. Some changes will require additional HERO tests. It is assumed that when the weapon system has reached the production phase, the design of all components and subsystems has progressed to a stage where a series of routine tests, evaluations, and analyses will be sufficient to verify that the weapon complies with requirements for precluding HERO problems, and can be certified for unrestricted use in the Fleet.

10-9. HERO TEST FOR WEAPON DESIGNERS.

MIL-STD-464 establishes the general requirements and acceptance criteria for precluding HERO problems associated with electrically initiated weapon systems components. Methods of implementing these requirements have been established and are presented in this HERO design guide.

The system evaluation tests performed prior to final acceptance of the weapon system by the Navy (see paragraph 10-3) cannot be conducted until all components have been fabricated and the complete weapon assembled. By the time this phase has been reached, the design is firm, and in many cases, production of the weapon has started. If the weapon fails to meet the HERO evaluation criteria, costly retrofits and redesign techniques may be required.

Previous Navy evaluation tests have demonstrated that little consideration was given to the HERO problem during the design stages of weapons presently used by the Fleet. Visual inspection of these systems would have been sufficient to detect such obvious deficiencies as long, unshielded umbilical cables and wires; plastic sections; and access doors that must be opened in the EME. There are, however, some serious design deficiencies that will not always be discovered by a visual inspection. Among the most important of these deficiencies are inadequate shielding and filtering.

To detect such deficiencies, to optimize design, and to implement effective quality control, it is imperative to apply qualitative testing techniques during the system=s developmental stages. As a result of an extensive research program, a series of such testing techniques has been developed and presented in MIL-STD-1377 (NAVY). These test methods include techniques for evaluating the shielding effectiveness of weapon enclosures and connectors, and for measuring filter effectiveness.

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APPENDIX A

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